GENTILE

BASIC THEORY AND APPLICATION OF TUNNEL DIODES

NOSTRAND

BASIC THEORY AND APPLICATION OF
TUNNEL DIODES

By Sylvester P. Gentile

A detailed easy-to-understand coverage of the physics of electron tunneling with a unique and simplified explanation of amplification and oscillation using negative resistance. Includes the operation of associated circuit components, and detailed analyses of practical circuits for a wide range of uses. Profusely illustrated.

For the technician and practicing engineer familiar with electron-tube theory and application, this book presents a unique, *simplified* version of the physics of the relatively complex concept of electron tunneling; this presentation applies not only to tunnel diodes but also to any comparable negative resistance device employed for amplification and oscillation. A knowledge of transistors, although desirable, is not required for an understanding of tunnel diodes as presented here, but for those who desire it, a brief review of the fundamentals of transistors is presented in Appendix C.

Initial chapters deal with basic theoretical explanations. The phenomenon of *electron tunneling* is given a detailed and thorough coverage in Chapter 2. This is followed by a unique presentation of the concept of "negative resistance" on a first order (non-mathematical) basis which should completely dispel the "mystery of negative resistance" usually resulting from a purely mathematical presentation.

Simplified versions of the physics of operation of associated circuit components such as hybrid junctions, ferrite isolators, Hall-effect gyrators and isolators, and backward diodes are also discussed. Also included are thorough explanations of a large number of practical tunnel diode circuits for use in computers, radar, microwave and UHF receivers and transmitters, as well as low frequency circuits.

For ease of comprehension the book is copiously illustrated with more than 200 line drawings and photographs.

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BASIC THEORY AND APPLICATION OF TUNNEL DIODES

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PREFACE

The aim of this book is to provide a coherent, sequential presentation of the theory and application of tunnel diodes consistent with the present state of the art. The level of treatment is the same as that given transistors in Department of the Army Technical Manual TM 11–690, "Basic Theory and Application of Transistors." That technical manual, available through the Superintendent of Documents, Washington, D. C., was prepared by this author with the assistance of several of his colleagues while employed at the Publications Engineering Department, United States Army Signal Materiel Support Agency, Fort Monmouth, New Jersey. As in the case of the transistor manual, this book is intended for the technician and the practicing engineer familiar with electron-tube theory and application. A knowledge of transistors, although desirable, is not required for an understanding of tunnel diodes as presented here. However, a brief review of the fundamentals of transistors is presented in Appendix C.

Chapters 2 and 3 are perhaps the most important chapters in the book. Chapter 2 provides the reader with a basic explanation of the internal conduction mechanism in heavily doped semiconductor diodes that results in the phenomenon of *electron tunneling*. Chapter 3 presents what the author believes to be a novel, but technically correct, explanation that will eliminate for most technicians and practicing engineers the *mystery* of *negative resistance* and its use for amplification and oscillation. This presentation applies not only to tunnel diodes but to any negative-resistance device employed for amplification and oscillation. After the contents of these two chapters are digested, the data presented in the remaining chapters should be more easily comprehended by the reader.

The author gratefully acknowledges the support and encouragement of Mr. Harry Barr, Chief, Surveillance Division, Publications Engineering Department, U. S. Army Signal Materiel Support Agency. Thanks are also due to Mr. Joseph Betzko for assistance in editing the manuscript and preparing the index; to Miss Helen Root, Miss Sandra Brewer, and Miss

Arlene Breglia for typing the manuscript; and to my son Arthur (age 13) for collating pages and numbering illustrations.

Acknowledgment is also made to the many engineers whose works and published articles (see Appendix A) have made this book possible, and to the General Electric Company and the Radio Corporation of America for supplying useful data.

It must be noted that this book was prepared by the author as a private individual; no approval of the contents, tacit or express, by the Department of the Army is claimed or granted.

SYLVESTER P. GENTILE

CONTENTS

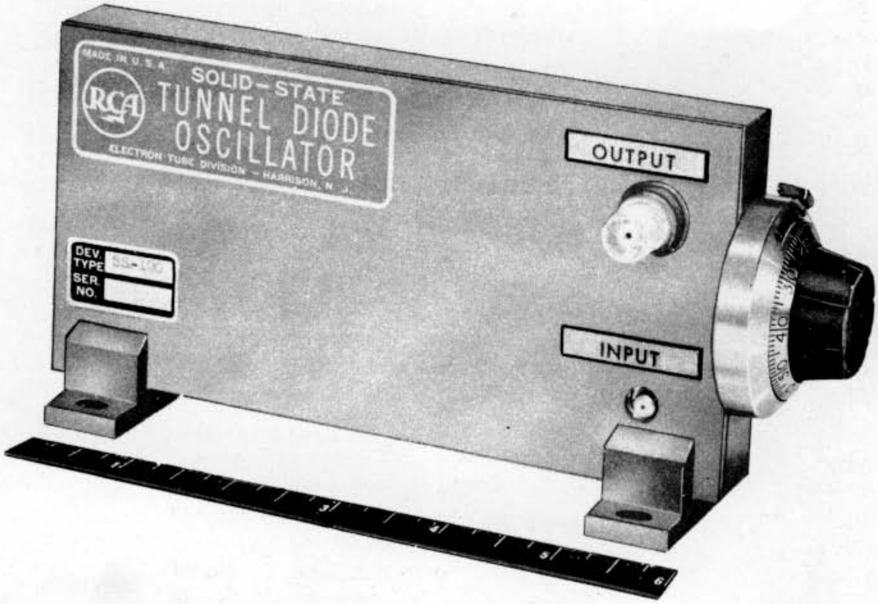


Fig. 1–1. Tunnel diode tunable (1050–1400 mc) oscillator (Courtesy Radio Corporation of America)

Chapter 1

INTRODUCTION

1–1. General

a. Fundamentally, the tunnel diode is a two-terminal semiconductor device that displays an ac (alternating current) *negative* resistance over a portion of its current-voltage curve. This phenomenon occurs if the tunnel diode is biased in the forward direction. It is this ac negative-resistance property of the tunnel diode that is exploited in a number of circuit arrangements to provide amplification, oscillation, switching, and memory functions extensively used in electronics.

b. Devices displaying ac negative resistance are not new to the field of electronics. This characteristic is found also in thyratrons, tetrode vacuum tubes, four-layer diodes, and others. However, because the tunnel diode is such a simple device, and a potentially inexpensive item, much engineering research has been directed toward increasing the number of applications. As a result, types of equipment using tunnel diodes are becoming more numerous and are being offered on the commercial market.

1–2. History of Semiconductor Devices

a. Crystal Rectifier. The first use of a crystal semiconductor as a rectifier (detector) was in the early days of radio. A crystal was clamped in a small cup or receptacle and a flexible wire (cat whisker) made light contact with the crystal. Tuning of the receiver was accomplished by operating the adjusting arm until the cat whisker was positioned on a spot of the crystal that resulted in a sound in the headset. Tuning the variable capacitor provided maximum signal in the headset.

b. Point-Contact Diode (Fig. 1–2). Point-contact diodes (germanium rectifiers) were used during World War II for radar and other high-frequency applications to replace electron-tube diodes. The point-contact diode has a very low shunt capacitance and does not require heater power; these properties provide a definite advantage over the electron-tube diode in high-frequency mixing and detecting applications. The point-contact

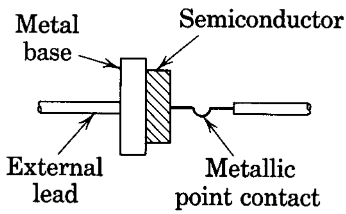


FIG. 1–2. Physical construction of point-contact diode

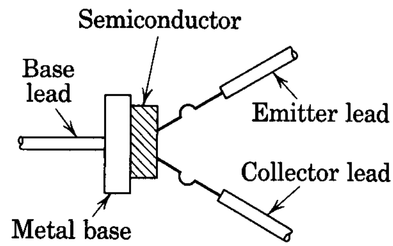


FIG. 1–3. Physical construction of point-contact transistor

diode is identical to the crystal rectifier (*a* above). The point-contact diode consists of a semiconductor, a metal base, and a metallic point contact. The connections to the point-contact diode are an external lead welded to the metallic point contact, and an external lead welded to the metal base.

c. Point-Contact Transistor (Fig.1–3). The development of the point-contact transistor was announced by the Bell Telephone Laboratories in 1948. The physicists credited with the invention were John Bardeen and Walter H. Brattain. The physical construction of the point-contact transistor is similar to that of the point-contact diode except that a third lead with a metallic point contact is placed near the other metallic point contact on the semiconductor. One lead is called an *emitter* lead; the other, a *collector* lead. When the two metallic points are properly biased with respect to the metal base, the point-contact transistor is capable of producing a power gain.

d. Junction Diode (Fig. 1–4). Rectification by semiconductor-to-semiconductor contact (a junction diode) was described in 1946 by W. H. Brattain of the Bell Telephone Laboratories. The rectifying junction diode is

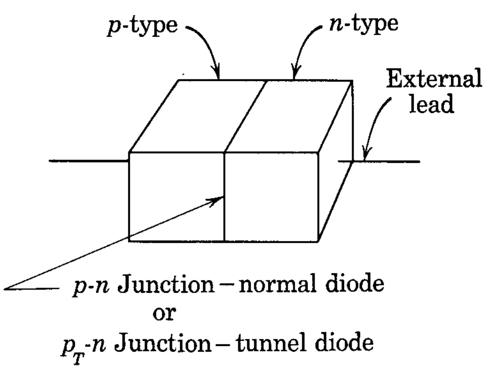
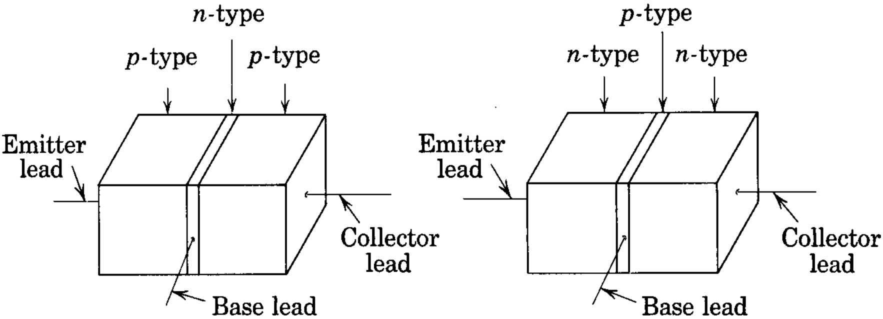


FIG. 1–4. Physical construction of rectifying junction diode, or tunnel diode

a single bar, slab, or wafer containing two dissimilar sections of semiconductor material. One section, because of its characteristics, is called a *p-type semiconductor;* the other, an *n-type* (par. 2–9 and 2–10). The connections to the rectifying junction diode consist of a lead to the *p*-type semiconductor and a lead to the *n*-type semiconductor. The rectifying junction diode can handle larger power than the point-contact diode but the rectifying junction diode has a larger shunt capacitance.

e. Junction Transistor. In 1949 W. Shockley of the Bell Telephone Laboratories published a theoretical analysis in which he predicted that a junction transistor consisting of area contacts (junctions) rather than point contacts, would be a practical device. Soon after, development of the junction transistor was announced. The junction transistor consists of two *p-n* junctions (Fig. 1–5A and B). Operation of the junction transistor is similar to that of the point-contact transistor. The junction transistor permits more accurate prediction of circuit performance, is less noisy, and is capable of handling more power than the point-contact transistor.

f. Tunnel Diode. The tunnel diode was invented in 1958 by a Japanese scientist, Leo Esaki. Thus, the tunnel diode is often referred to as the Esaki diode. The external appearance and structure of the tunnel diode is identical to the normal rectifying junction diode (Fig. 1–4). The difference between the two is in the amount of donor and acceptor impurities (Ch. 2) added to the semiconductor materials. The impurities added to the tunnel diode semiconductor materials are approximately 1000 times greater than those used in semiconductor materials for the rectifying junction diode. The junction formed in the rectifying diode is referred to as a *p-n* junction; in the tunnel diode the junction is referred to as a p_T-*n* junc-



A. *p-n-p* (semiconductor) junction transistor B. *n-p-n* (semiconductor) junction transistor

FIG. 1–5. Physical construction of *p-n-p* and *n-p-n* junction transistors

tion. It is the heavier doping (adding of impurities) of the tunnel diode semiconductor materials which causes the phenomenon of ac negative resistance (par. 3–7) that permits the use of the tunnel diode as an amplifier, oscillator, switch, etc.

1–3. Tunnel Diode Functions

a. Amplification. The tunnel diode may be used as a current, voltage, or power amplifier. For instance, a stronger signal current may be obtained

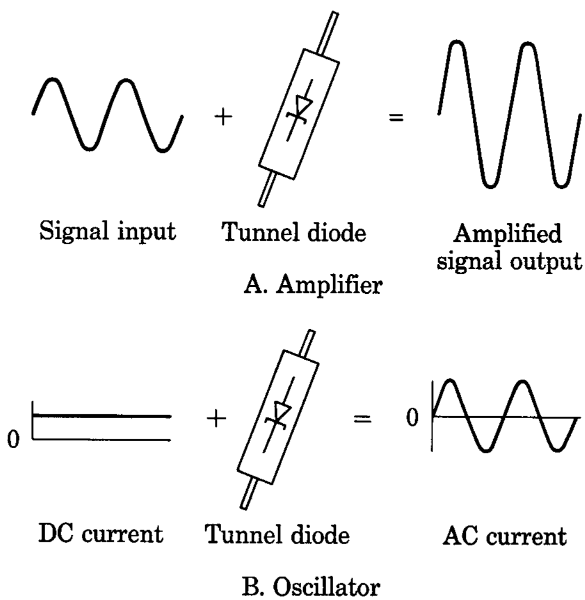


Fig. 1–6. Tunnel diode used as amplifier or oscillator

from a tunnel diode (Fig. 1–6A) than is delivered by the signal source. Various circuit arrangements provide for various amounts of signal amplification, depending on the power-handling capacity of the particular tunnel diode.

b. Oscillation. The tunnel diode may be used to convert direct-current energy into alternating-current energy; i.e., it may be used as an oscillator. When functioning in this manner, the tunnel diode shifts energy from a dc source and, in conjunction with a suitable timing (tank) circuit arrangement, generates an ac signal (Fig. 1–6B).

c. Modulation and Demodulation. The tunnel diode used in various circuit arrangements can provide amplitude modulation (variation in ampli-

tude of an RF signal) (Fig. 1–7A) or frequency modulation (variation in frequency of an RF signal) (Fig. 1–8A). Demodulation (detection) of amplitude-modulated signals (Fig. 1–7B) or frequency-modulated signals (Fig. 1–8B) may be accomplished with tunnel diodes. These circuits are well-suited for miniature transmitters intended for short-range applications.

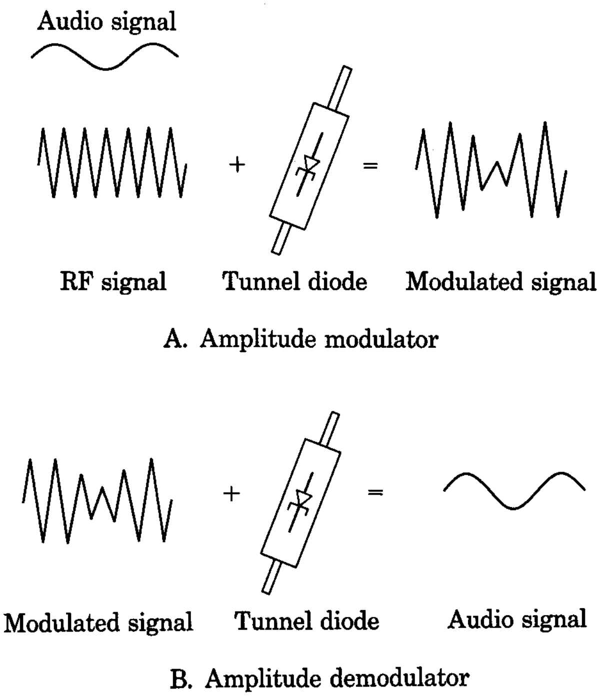


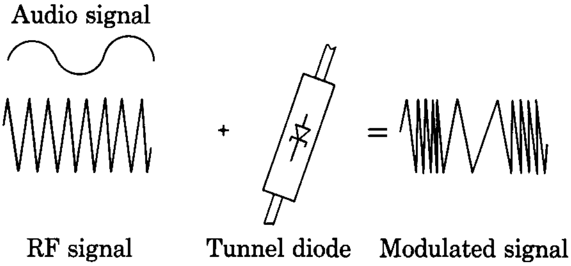
FIG. 1–7. Tunnel diode used as AM modulator or AM demodulator

d. Miscellaneous. The tunnel diode may also be used to modify the shape of signal waveforms. Waveform shaping is vital in various types of radar, teletypewriter, computer, and television circuits. Figure 1–9 indicates the use of the tunnel diode in transforming a sinewave into a square wave.

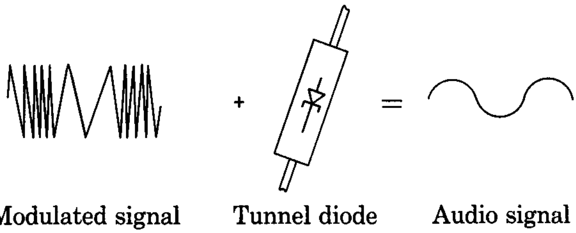
1–4. Comparison of Transistors and Tunnel Diodes

a. The transistor and the tunnel diode have common advantages over electron tubes. These are:

1. Greater power efficiency because no heater element is involved.
2. No warm-up time required.



A. Frequency modulator



B. Frequency demodulator

FIG. 1–8. Tunnel diode used as FM modulator or FM demodulator

3. Long life expectancy because of their physical ruggedness. Each will withstand centrifugal force, gravity, and impact tests that would completely shatter an electron tube.

4. Operation at low voltages that permit the use of smaller circuit components.

5. Extremely small in size compared to even miniature electron tubes. This permits ease of construction in microminiature modules especially desirable for use in portable equipment, aircraft, missiles, drones (unmanned aircraft), and satellites.

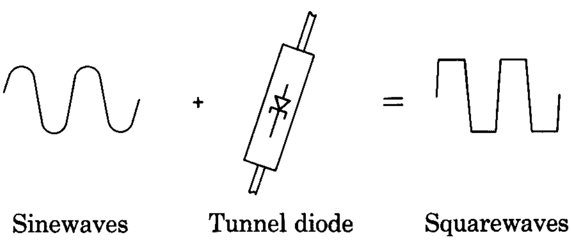


FIG. 1–9. Tunnel diode used to square waveform

b. The tunnel diode has the following advantages over transistors:

1. Much higher switching speeds. Present tunnel diodes can switch in one nanosecond (10^{-9} second or one billionth of a second). The tunneling action that occurs in a tunnel diode takes place theoretically at nearly the speed of light. Switching speed is of particular use in high-speed computers. Because of this fact it is expected that tunnel diodes will be extensively applied in this field. The transistor is limited in switching speed because of the long transit time of current carriers through its base.

2. The tunnel diode can operate through wide temperature ranges without appreciable changes in its tunneling characteristics. This property is due to the very heavy doping of the semiconductor material. Extremes of heat from absolute zero to 400° C (for gallium arsenide tunnel diodes) do not generate enough electron-hole pairs and excess electrons (par. 2–18) to increase markedly in percentage the original number of these carriers contributed by the doping materials. In the transistor, however, light doping is used and an operating temperature of 100° C materially affects its characteristics.

3. The tunnel diode can withstand relatively much larger doses of nuclear radiation compared to the transistor before its characteristics are changed. Again the reason is the heavy doping of semiconductor material (2 above).

4. Because the tunnel diode is a two-terminal device, there is no necessity to connect a third lead to an extremely small area as in the case of the transistor. This, plus the fact that only two regions of semiconductor material are required instead of three regions, as in the transistor, should result in a relatively inexpensive device.

5. Tunnel diode amplifiers can operate effectively even at microwave frequencies, producing wide bandwidth and high gain at low noise figures. Transistors have not competed with masers and parametric amplifiers at microwave frequencies. The high junction capacitances in the transistor and the long transit time of current carriers through the base region preclude use of the transistor at these frequencies. Tunnel diodes will be much lower in cost than masers and parametric amplifiers.

6. Heavy overload current of short duration will not permanently damage the tunnel diode. Actually, it is difficult to produce sufficient power in its junction to damage it.

7. Tunnel diode properties and characteristics are relatively unchanged by moisture and atmospheric gases. This fact eliminates the need for hermetic sealing that is required for transistors. The tunnel diode can be produced in very small, lightweight, encapsulated units requiring less space than transistors.

c. The main disadvantages of the tunnel diode are:

1. Gain for amplifiers and oscillators can be obtained only by operating over the ac negative-resistance characteristic of the device.
2. The tunnel diode has only two terminals that must be used for both input and output. Obviously the tunnel diode is not a unidirectional device. In simple circuit arrangements the output directly affects the input. Cascading amplifier stages becomes a major design difficulty. However, suitable unidirectional circuit arrangements can overcome this difficulty (Ch. 5).
3. The tunnel diode is a very low-voltage signal device. The available voltage swing of germanium units is less than ½ volt; gallium arsenide units provide less than 1 volt. Peak currents, however, can range from 10 μa to more than 10 amp.

1–5. Tunnel Diode Material

Materials, such as copper, silver, gold, and iron, which provide a good path for electron flow with little opposition (resistance), are referred to as *conductors* (par. 2–5*a*). Materials such as carbon in diamond form, germanium, and silicon, which provide a path for electron flow but offer moderate opposition, are referred to as semiconductors (par. 2–5*c*). Materials such as rubber, porcelain, and glass, which offer great opposition and do not provide a path for electron flow, are referred to as insulators (par. 2–5*b*). Tunnel diodes are composed of semiconductor materials such as germanium or silicon.

1–6. Summary

a. The ac negative-resistance characteristic of the tunnel diode may be employed in various circuit arrangements to perform functions normally performed by transistors and electron tubes.

b. The tunnel diode is made of a heavily doped semiconductor material, such as germanium or silicon.

c. The first crystal semiconductor was used as a rectifier (detector) in the early days of radio.

d. The point-contact transistor was invented in 1948.

e. The junction transistor was invented in 1949.

f. The tunnel diode was invented in 1958.

g. The tunnel diode may be used in circuits, such as amplifiers, oscillators, modulators, and demodulators.

h. The tunnel diode is smaller and more rugged than the electron tube. In addition, the power efficiency of the tunnel diode is greater than that of the electron tube.

i. Tunnel diodes are especially adaptable to the miniaturization of electronic equipment. Even the associated circuit components of tunnel diodes, such as capacitors, resistors, batteries, and transformers, may be made smaller than the corresponding components used in electron-tube circuits.

j. The advantages of the tunnel diode over the transistor are: faster switching; higher tolerance to high temperatures, nuclear radiation, and momentary current overloads; amplification and oscillation in the kilomegacycle region; high tolerance to moisture and atmospheric gases; lighter and less expensive to manufacture.

k. The main disadvantages of the tunnel diode are that it offers a small usable signal-voltage swing and it is not a unidirectional device.

Chapter 2

FUNDAMENTAL THEORY OF TUNNEL DIODES

SECTION I. STRUCTURE OF MATTER

2–1. General

a. A knowledge of the theory of the structure of matter is required for an understanding of the theory of the internal conduction mechanism that occurs in the tunnel diode.

b. Tunnel diodes are constructed from semiconductor materials. A comparison of the properties of conductors, semiconductors, and insulators is given in paragraph 2–5. Detailed properties and characteristics of semiconductor materials used in transistors and tunnel diodes are covered in paragraphs 2–6 through 2–12.

2–2. Matter, General

Matter is any substance that has weight (mass) and occupies space. Examples of matter are: air, water, plants, and metals. As these examples indicate, matter may be found in the gaseous, liquid, or solid state. Matter is found in nature as elements (*a* below), or compounds (*b* below). The elements and compounds are made up of molecules (*c* below), atoms (*d* below), and subatomic particles (*e* below).

a. Element. Matter consists of one or more basic materials which are called elements. Scientists have definite proof that at this time 102 elements exist. An element is defined as a substance that can be neither decomposed (broken up into a number of substances) by ordinary chemical changes nor made by chemical union of a number of substances. Copper, iron, aluminum, and gold are examples of metallic elements; oxygen, hydrogen, and sulfur are nonmetallic elements.

b. Compound. A substance containing more than one element and usually having properties different from those of its elements is called a *compound*. For example, water is made up of hydrogen and oxygen. Therefore, water is a compound.

c. Molecule. A *molecule* is defined as the smallest particle of matter that can exist by itself and still retain the properties of the original substance. If a drop of water, a compound, is divided until the smallest possible particle is obtained and is still water, that particle is known as a molecule. An idea of the size of a molecule may be obtained by imagining that a stone is first broken into two pieces, that the two pieces are then broken into four pieces, and that this process is continued. The smallest particle of stone which could be obtained by this process would be a molecule. Actually, it is impossible to crush a stone into its molecules; we can only crush it into dust. One small particle of dust is composed of thousands of molecules.

d. Atom. An atom is defined as the smallest part of an element that can take part in ordinary chemical changes. For simplicity, the atom may be considered to be the smallest particle that retains its identity as part of the element from which it is divided. Since there are approximately 102 known elements, there must be 102 different atoms, or a *different* atom for each element. All substances are made of one or more of these atoms.

e. Subatomic Particles. The atom itself can be subdivided into still smaller, or subatomic, particles. The nature of these subatomic particles is covered in paragraph 2–4.

2–3. Structure of Atom

Figure 2–1 shows the structure of matter. Parts A and B of the illustration are real; parts C through F are imaginary but are based on extensive laboratory and theoretical data.

a. Figure 2–1A shows the metal as it appears to the unaided eye.

b. In Fig. 2–1B the magnification is 100 diameters. Note that the metal has a crystalline structure. The crystals are small, nonuniform in shape, and irregularly arranged. This arrangement of crystals is referred to as a *polycrystalline structure.* All metals reveal a polycrystalline structure when seen through a microscope. The properties and characteristics of polycrystalline materials are quite different from the properties and characteristics of *single* crystal materials. Germanium and silicon, when processed for use in semiconductor devices, are *single crystal* materials (par. 2–7).

c. Magnification to 100,000 diameters (Fig. 2–1C) gives evidence of the presence of individual atoms or subatomic particles.

d. In Fig. 2–1D magnification is 10 million diameters. At this magnification the metal atoms appear as identical dots in straight rows.

e. At a magnification of 100 million diameters a single metal atom fills the entire area (Fig. 2–1E). This single atom resembles the solar system with a central body, called a nucleus, about which a number of smaller particles (electrons) move in outer orbits. Each of the electrons in this atom has a charge of electricity identical with the charge on any of the

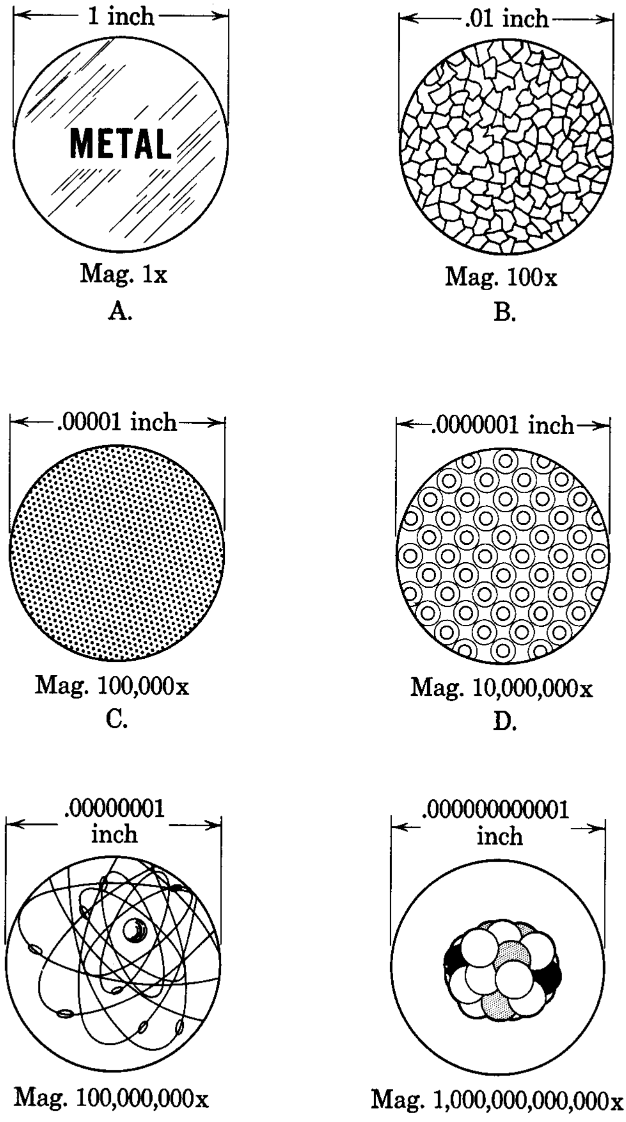


FIG. 2-1. Structure of a metal element

other electrons. The charge associated with an electron, the elemental charge, is the smallest electrical charge. The charge on the electron is arbitrarily designated a *negative charge*. For each negative electron that orbits about the nucleus, there is a positive proton in the nucleus, so that the atom is electrically neutral.

f. Increased magnification (Fig. 2–1F) shows that the enlarged nucleus contains two kinds of particles. The positively charged particles are called *protons*. The uncharged particles are called *neutrons*.

2–4. Electrons, Protons, and Neutrons

It has been established that the metal atom consists of a positively charged nucleus with negatively charged electrons that orbit around the nucleus.

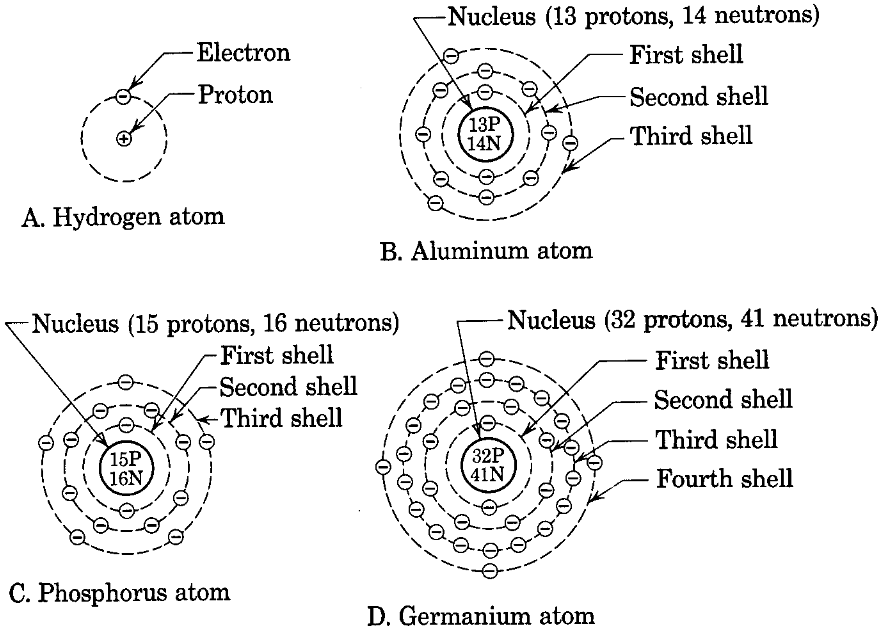


FIG. 2–2. Structure of atoms

Actually the atoms of *all* elements (oxygen, silver, hydrogen, etc.) contain a central nucleus and orbiting electrons.

a. Examples of Atomic Structure. 1. The atomic structure of the hydrogen atom (Fig. 2–2A) is the simplest of all atoms. It contains one electron orbiting around the nucleus which consists of one proton. The negative charge on the electron equals the positive charge on the proton and the atom is electrically neutral.

2. The nucleus of the aluminum atom (Fig. 2–2B) contains 14 neutrons and 13 protons. The positive charges of the 13 protons balance the negative charges of the 13 electrons and the entire atom is neutral. Note that the outermost shell has three electrons. The importance of these three electrons is explained in paragraph 2–10.

3. The phosphorus atom (Fig. 2–2C), a more complex structure, has 15 orbital electrons in three separate rings or *shells*. In this atom the outermost ring has five electrons. The importance of these five electrons is explained in paragraph 2–9.

4. The germanium atom (Fig. 2–2D), an even more complex atom, has 32 protons and 41 neutrons in the nucleus. The 32 orbital electrons revolve in four separate shells, with four electrons revolving in the outer incomplete shell. Paragraph 2–7 explains the importance of this arrangement.

b. Building Blocks and Their Characteristics. The difference in the various elements and their characteristics is in the *number* and *arrangement* of the electrons, protons, and neutrons of which each atom is composed. All electrons are basically identical regardless of the atom, and therefore the element, of which they are a part. The same can be said of all protons and all neutrons. Since matter is composed of atoms of positively charged protons, negatively charged electrons, and uncharged neutrons, these charged and uncharged particles are the fundamental building blocks of all matter. The charge on the negative electron or the positive proton is considered the elemental unit of electrical charge. Because the elemental unit is too small a quantity of electricity for practical purposes, a larger unit called the coulomb is commonly used. One coulomb of electricity represents 6.28 million, million, million (6.28×10^{18}) electrons. Although they have equal and opposite quantities of charge, a proton's mass is 1850 times greater than the mass of the electron. The neutron is equal in diameter and mass to the proton. Relatively great distances exist between the electrons and the protons of an atom. A copper one-cent piece magnified to the size of the earth's path around the sun (approximately 584,000,000 miles) would show electrons the size of baseballs spaced about 3 miles apart.

2–5. Conductors, Semiconductors, and Insulators (Fig. 2–3)

In the field of electricity all materials are placed in three main categories: conductors, semiconductors, or insulators. The category into which a material is placed depends on its ability to conduct electricity. This, in turn, depends on its atomic structure.

a. Conductors. A good conductor is a material that has a *large number of loosely held electrons*. All metals conduct electricity, but some are better conductors than others. A one-centimeter cube (each edge measures one centimeter) of silver, copper, or aluminum has a resistance of less than three millionths of an ohm. Silver is a better conductor than copper, but copper is used more extensively because it is less expensive. Aluminum is used as a conductor where weight is a major factor, for example, on long-span high-tension lines.

b. Insulators. A material classified as an insulator has an atomic structure that does not permit the movement of electrons from atom to atom. An insulator (or dielectric) is a material that has few loosely held electrons. Actually there are no perfect insulators. However, such materials as glass,

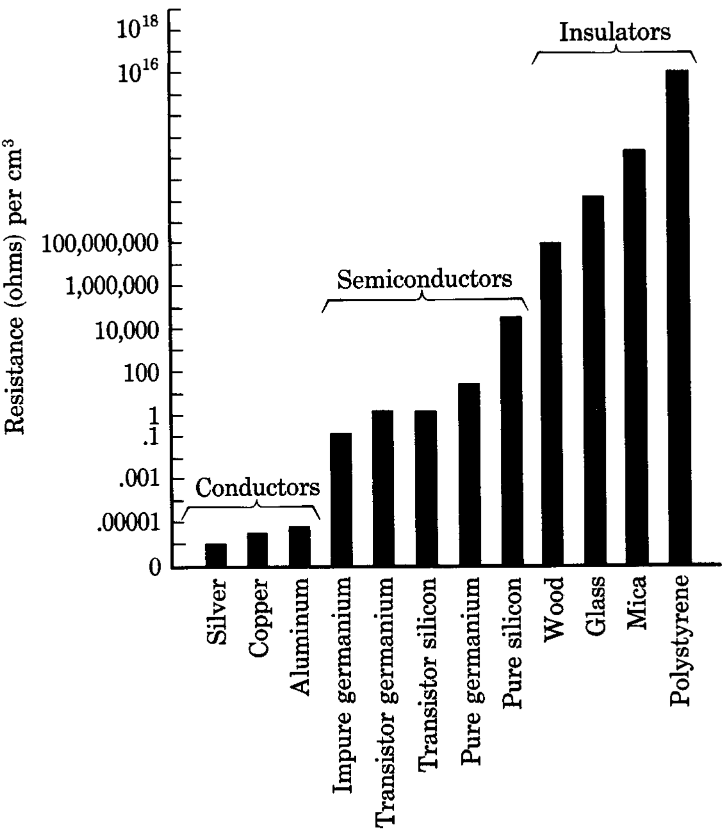


FIG. 2-3. Chart of resistance in ohms for a centimeter cube of conductors, semiconductors, and insulators

rubber, mica, and polystyrene have a resistance of several millions of ohms per centimeter cube. For most practical purposes these are considered nonconducting materials.

c. Semiconductors. Semiconductors are materials which are neither good conductors nor good insulators. Germanium and silicon fall into this category. At room temperature pure germanium has a resistance of 60 ohms per centimeter cube. Its resistance is several million times greater than that of copper and several million times less than mica. Pure silicon is a

worse conductor, having a resistance of 60,000 ohms per centimeter cube at room temperature. Both impure germanium and impure silicon derived from related compounds have a resistance of 0.2 and 0.4 ohm per centimeter cube, depending on the type and amount of impurities present. Germanium and silicon, for use in transistors and rectifying diodes, have carefully controlled amounts of impurities added (par. 2–8 and 2–10), and each has a resistance of 2 ohms per centimeter cube at room temperature. This resistance decreases rapidly as the temperature rises. The resistance of semiconductor materials used for tunnel diodes is 1000 times less than the resistance of semiconductor materials used for transistors and rectifying diodes. Note that while only germanium and silicon as semiconductors are discussed, there are many other semiconductors.

SECTION II. IMPURITIES AND CURRENT CARRIERS IN CRYSTALS

2–6. Crystals, General

a. General. Most solids, except those exhibiting a biological structure of cells, such as leaves, branches, and bone, reveal a crystal structure when studied under a microscope. Many substances, such as rocks and metals which are not usually considered crystalline, reveal a specific crystal pattern when studied under a microscope.

b. External Characteristics. The most commonly known characteristics of crystals are their angles and their planes. Snow crystals, for instance, although formed in an infinite number of geometric patterns, contain only 60° angles. Some materials, such as common salt (sodium chloride), form cubes; other materials form long needles, rhomboids, or variations of hexagonal or rectangular structures. Each material has a characteristic form.

c. Internal Structure. X rays have been used to investigate the internal structure of crystals. The wavelengths of X rays approximate the distance between the atoms or molecules of crystals. When X rays are beamed through a crystal, the rays are deflected and distributed in accordance with the specific arrangement of the atoms or molecules of the crystal. When the resultant deflected rays are photographed, the photograph invariably shows a specific pattern depending upon the substance of the crystal. With the pattern indicated on the photograph, and through complex mathematical analyses, scientists have been able to construct models of the internal structure of a given crystal. These analyses have indicated that the atoms of crystals are arranged in specific patterns; that one atom is not closely related with another atom only, but rather is related equally to a number of adjacent, equidistant atoms. The specific arrangement of atoms depends on the size and number of atoms present and on the electrical forces between them. The physical, electrical, optical, and mechanical characteristics of the

crystal depend on the forces between the atoms. The crystals discussed in paragraphs 2–7 through 2–8 are semiconductor crystals, such as germanium and silicon. These materials are used in making semiconductor devices, e.g., tunnel diodes and transistors. The most important common characteristics among these materials is that each of their atoms has four electrons in the outer shell. The discussion therefore applies to any of these semiconductor materials.

2–7. Pure Semiconductor Crystal

a. General. Figure 2–4 shows a pure semiconductor crystal. Each sphere represents a semiconductor atom less the four electrons (valence or outer-orbit electrons) that are in the outer (fourth) incomplete shell of the atom (Fig. 2–2D). The sphere (Fig. 2–4) contains the nucleus of the atom, and all the tightly bound electrons that orbit around the nucleus. Since the atom of any element is electrically neutral, the sphere has a net positive charge of four. Throughout this text, the sphere will be referred to as the *semiconductor core.*

b. Single Crystal Structure. The dashed lines in Fig. 2–4 form two cubes. Note that the four semiconductor cores between the two cubes are shared equally by the cubes. If the illustration were to be extended in all directions, the sharing of the corner semiconductor cores would be extended to all adjacent cubes. This repeated, uniform, *cubical structure* constitutes a *single semiconductor crystal.* The properties and characteristics of *single crystal* materials, such as germanium and silicon (as prepared for use in semiconductor devices), are quite different from the properties and characteristics of *polycrystalline materials,* e.g., copper and aluminum (par. 2–3*b*). The term crystal used throughout this text will refer to *single crystal material* only.

c. Lattice Structure. 1. It has been established that electrons rotate constantly in relatively fixed orbits about the nucleus. In a crystal, the rotation of one valence electron of a given atom is coordinated with the rotation of one valence electron of an adjacent atom. The coordinated rotation of two valence electrons (one from each of two adjacent atoms) results in the formation of an *electron-pair bond.* The electron-pair bonds shown diagrammatically in Fig. 2–4 are also referred to as *valence bonds.* The electron-pair bonds cause the cores to be attracted toward each other. The positive charges on the cores cause the cores to repel each other. When a balance of the forces of attraction and repulsion is obtained, the crystal is said to be in a state of equilibrium.

2. Each semiconductor core is equidistant from four adjacent semiconductor cores. Note that each core is interconnected with adjacent cores by four electron-pair bonds. This condition exists since each semiconductor

atom contains four valence electrons in its outer shell. This arrangement of semiconductor cores and electron-pair bonds is referred to as a *lattice*. To avoid congestion, Fig. 2–4 shows only a few of the electron-pair bonds.

d. Conductivity. The valence electrons of such good conductors as copper or aluminum are loosely bound to the nucleus of the atom, and they move quite readily through the conductor under the influence of an electric or magnetic field. Valence electrons which form part of an electron-pair bond,

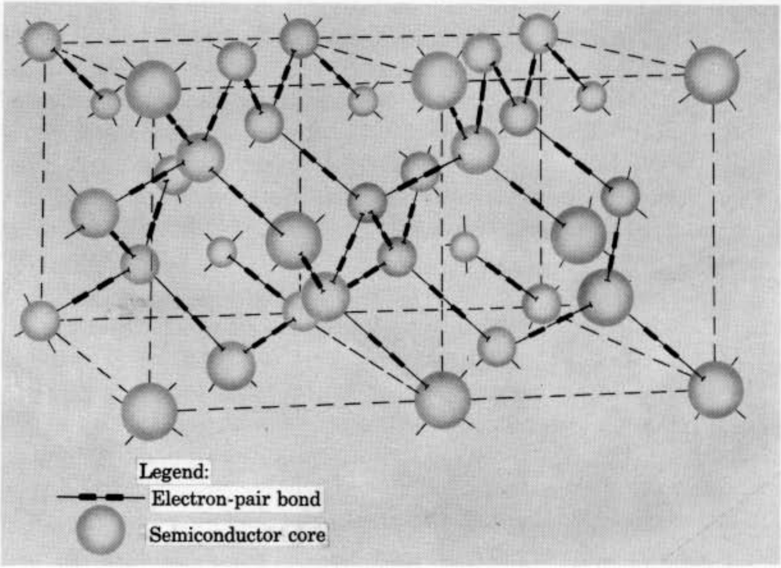


FIG. 2–4. Pure semiconductor crystal, lattice structure

however, are bound in the electron-pair bond and are not free to take part in conduction. Crystalline materials, such as germanium and silicon, the valence electrons of which are bound, are poor conductors (Fig. 2–3) under normal conditions. Only if the material is subjected to high temperatures or strong radiation will the electron-pair bonds separate and partial electrical conduction occur.

2–8. Impurities

a. General. It is possible for the atoms of substances having more or less than four valence electrons to join the crystal lattice structure of the semiconductor material. These substances whether found in the semiconductor material in its natural state, or added intentionally during the processing of the semiconductor material for use in tunnel diodes or transistors, are referred to as *impurities*.

b. Donor and Acceptor Impurities. Two groups of substances exhibit the important characteristic of joining the lattice structure of semiconductor materials. The substances in one group are called *donors;* in the second group, they are called *acceptors.*

1. The atoms of substances classified as donors have five valence electrons in the outer incomplete shell. Some of the substances that have been used as donors are phosphorus (Fig. 2–2C), arsenic, and antimony. The characteristics and properties of semiconductor materials containing donor atoms are discussed in paragraph 2–9.

2. The atoms of substances classified as acceptors have three valence electrons in the outer incomplete shell. Some of the substances that have been used as acceptors are aluminum (Fig. 2–2B), gallium, boron, and indium. The characteristics and properties of semiconductor materials containing acceptor atoms are covered in paragraph 2–10.

2–9. *n*-Type Semiconductor

a. Figure 2–5 shows a semiconductor crystal in which one of the semiconductor atoms has been replaced by a donor impurity (par. 2–8*b*1). The dark sphere in the illustration represents the nucleus of the donor atom and all the tightly bound electrons that orbit around the nucleus. The valence electrons are not included in the sphere. The donor impurity contains five valence electrons. Note that four of the valence electrons of the

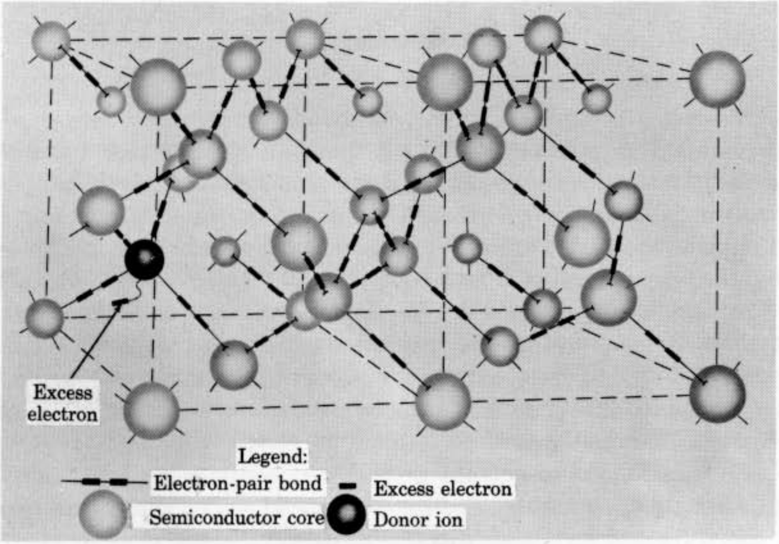


FIG. 2–5. Semiconductor crystal with donor atom

donor form electron-pair bonds with electrons of four neighboring semiconductor atoms. The electrons of both the semiconductor and the donor atoms that enter into electron-pair bonds form a very stable structure and are not readily removed from the bonds.

b. The fifth valence electron of the donor cannot form an electron-pair bond since there are no adjacent electrons available. This electron becomes an excess electron. The donor nucleus has a very weak influence over the

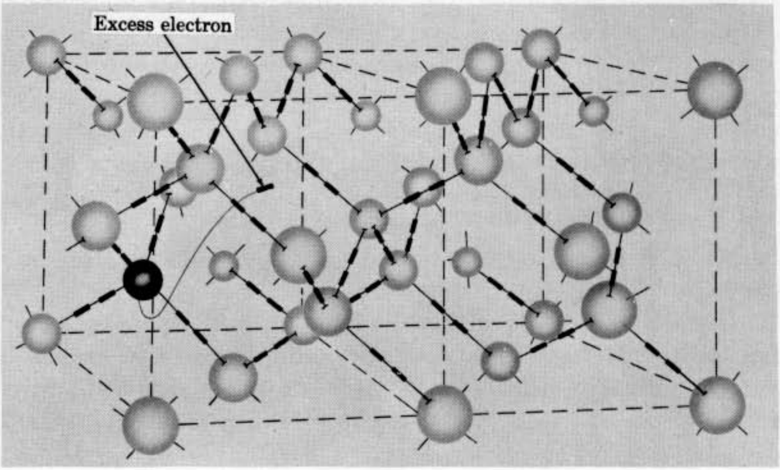


FIG. 2–6. Semiconductor crystal with donor atom showing movement of excess electron

excess electron. Actually, only one one-hundred-seventieth ($\frac{1}{170}$) of the energy required to remove an electron from an electron-pair bond is required to remove the excess electron from the donor. At normal room temperature (70° F), enough thermal (heat) energy is present to cause the excess electron to break away from the donor and wander through the space between the crystal lattices (Fig. 2–6).

c. When the excess electron leaves (ionizes from, or is donated by) the donor atom, the donor atom then possesses a positive charge equivalent to the negative charge of one electron. An atom that loses or gains an electron is called an *ion*. For that reason, the spheres that represent the donor (Figs. 2–5 and 2–6) are called donor ions.

d. Note that the semiconductor crystal that contains a donor ion (positive) also contains an excess electron (negative). The crystal taken as a whole therefore is electrically neutral; i.e., the crystal possesses a net charge of zero.

e. Semiconductors containing donor impurities are referred to as *n-type* semiconductors. The letter *n* refers to the *negative* charge of the excess electron.

2–10. *p*-Type Semiconductor

a. Figure 2–7 shows a semiconductor crystal in which one of the semiconductor atoms has been replaced by an acceptor impurity (par. 2–8*b*2).

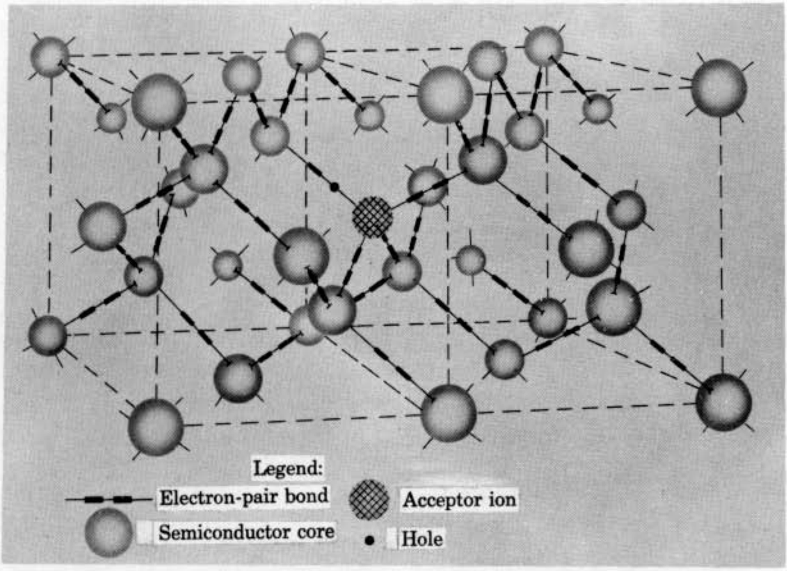


FIG. 2–7. Semiconductor crystal with acceptor atom

The cross-hatched sphere contains the nucleus of the acceptor atom and all the tightly bound electrons that orbit around the nucleus; the valence electrons are not included in the sphere. The acceptor impurity contains three valence electrons. Note that the three valence electrons of the acceptor form electron-pair bonds with electrons of the neighboring semiconductor atoms.

b. One valence electron of the fourth neighboring semiconductor atom cannot form an electron-pair bond because the acceptor has only three valence electrons. In this condition, an electron-hole arrangement exists. The position that would normally be filled with an electron is designated as a *hole*.

c. It is possible for an electron from an adjacent electron-pair bond to absorb enough thermal or electrical energy to break its bond (Figs. 2–8 and 2–9) and fill in the hole in the original electron-hole arrangement (Fig. 2–7).

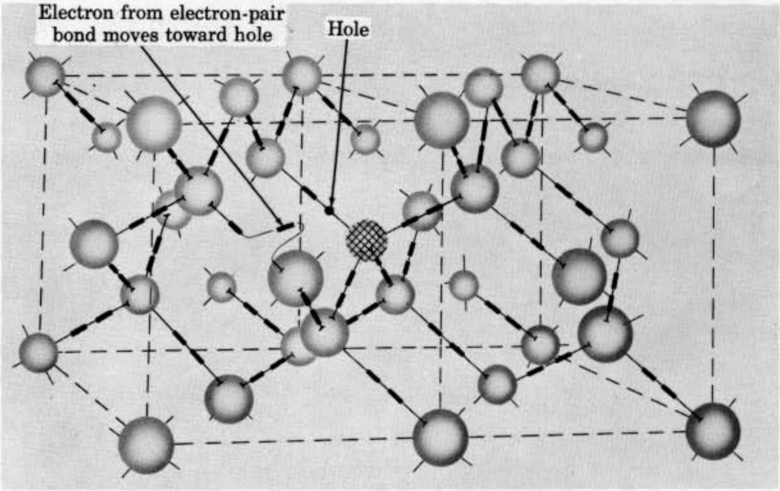


FIG. 2–8. Semiconductor crystal with acceptor atom, showing electron from electron-pair bond moving toward hole

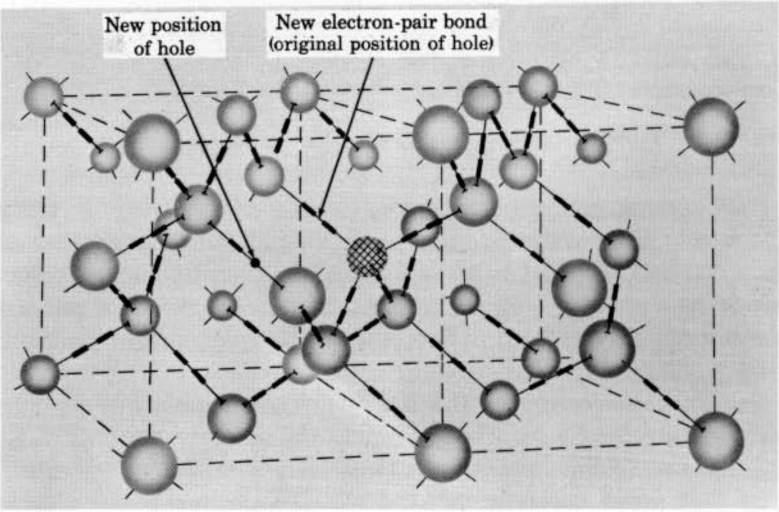


FIG. 2–9. Semiconductor crystal with acceptor atom showing new position of hole

Note that the hole has moved to a new position in Fig. 2–9. When the hole moves to the new position, two important changes take place.

1. The first change is that the acceptor atom has been ionized; i.e., the acceptor has acquired (or *accepted*) an electron and is now an ion. A negative charge exists in the immediate vicinity of the acceptor.

2. The second change is that the semiconductor atom, which requires four valence electrons, is left with only three valence electrons. The semiconductor atom, lacking an electron, has a net positive charge equivalent to the negative charge of the electron. Because of the crystal structure of the semiconductor, the positive charge of the semiconductor atom is not diffused or scattered, but is concentrated in the hole in the electron-hole arrangement. Furthermore, laboratory experiments have shown that the positive hole moves within the crystal in the same manner that an excess electron moves within the crystal. The concept of holes is very important in understanding the operation of semiconductor devices; the properties and characteristics of holes are discussed more thoroughly in paragraph 2–12.

d. Note that the semiconductor crystal which contains an acceptor ion (negative) also contains a hole (positive). The crystal taken as a whole therefore possesses a net charge of zero.

e. Semiconductor materials containing acceptor impurities are referred to as *p-type semiconductors.* The letter *p* refers to the *positive* charge of the hole.

2–11. Movement of Hole

Figure 2–10 is a two-dimensional representation of the mechanism involved in the movement of a hole through a crystal. In Fig. 2–10A, the hole is in the upper left-hand corner. An electron from an adjacent electron-pair bond moves to the position of the hole. The hole (Fig. 2–10B) is now midway between top and bottom of the crystal and slightly to the right of its original position. This process is repeated in Figs. 2–10C and D, until the hole is at the right-hand side of the crystal (Fig. 2–10E). The complete path of the hole through the crystal is shown in Fig. 2–10F.

2–12. Holes, Properties and Characteristics

For an understanding of the theory of semiconductor devices, it is convenient for the reader to think of the hole as a specific particle. Holes in motion, like electrons in motion, constitute an electrical current (par. 2–11). There are differences, however, which must be kept in mind:

a. The hole can exist only in a semiconductor material, such as germanium or silicon. This is because the hole depends for its existence on a specific arrangement of electrons (electron-pair bonds) and atoms as is found in

crystal substances (par. 2–10). Holes *do not* exist in such conductors as copper and aluminum.

b. The hole is deflected by electric and magnetic fields in the same manner that electrons are deflected. Because the hole possesses a charge equal and opposite to that of the electron, the direction of deflection of the hole is

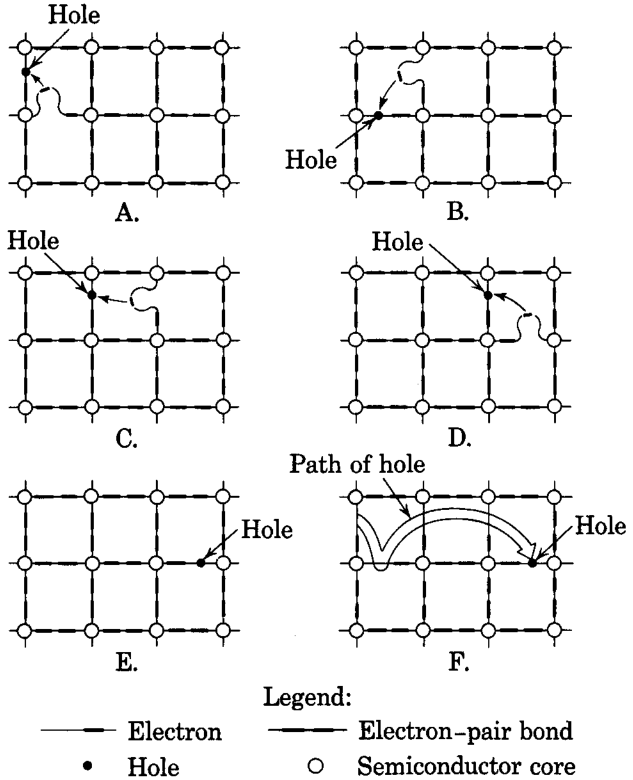


FIG. 2–10. Movement of hole through crystal

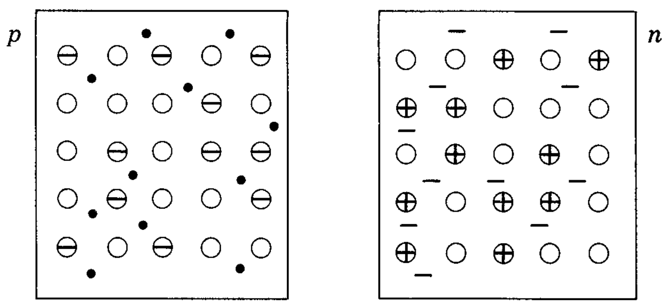
opposite to that of the electron. In an electric field, for instance, the electron moves toward the positive potential; the hole moves toward the negative potential.

c. In the field of electronics, the electron is considered indestructable. When a hole is filled by an electron from an adjacent electron-pair bond, the hole is considered as having moved from one position to another (Figs. 2–8 and 2–9). When a hole is filled by a free or excess electron, the hole no longer exists. This statement is supported by the fact that a semiconductor material containing an equal number of donor and acceptor atoms has none of the properties of the *p*-type or *n*-type semiconductor.

SECTION III. *p-n* JUNCTIONS

2–13. General

When an *n*-type semiconductor (par. 2–9) and a *p*-type semiconductor (par. 2–10) are joined in the same crystal, an unusual but important phenomenon occurs at the surface where contact is made between the two types of semiconductor. The contact surface is referred to as a *p-n junction.*



Legend:

○ Semiconductor core

⊖ Acceptor ion (negative)

⊕ Donor ion (positive)

• Hole (positive)

— Excess electron (negative)

FIG. 2–11. Separated sections of *p*-type and *n*-type semiconductor

The phenomenon that occurs at *p-n* junctions permits the use of semiconductors, such as germanium and silicon, in circuits normally employing electron tubes. The detailed theory of *p-n* junctions is covered in paragraphs 2–14 through 2–18.

2–14. *p-n* Junction, General

a. Figure 2–11 shows a section of *p*-type semiconductor and a section of *n*-type semiconductor.

1. For clarity, the electron-pair bonds are not shown; only the holes, the excess electrons, the semiconductor cores, and the donor and acceptor ions are represented.
2. For discussion purposes, Fig. 2–11 shows a large number of acceptor ions in the *p*-type and a large number of donor ions in the *n*-type germanium.

In practice, however, semiconductor materials, such as those used for transistors and rectifying diodes, contain approximately one impurity atom per 10 million germanium atoms.

3. If one could actually look inside the bulk semiconductor material, one would see the semiconductor cores and the impurity ions vibrating within their lattice positions because of thermal energy. However, the cores and

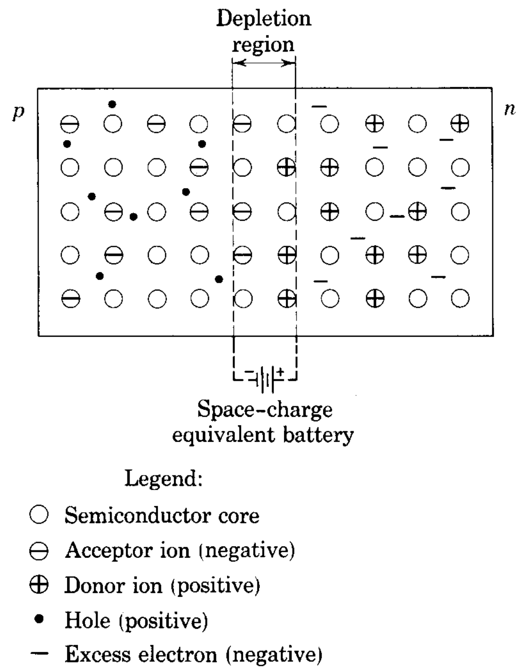


FIG. 2–12. Joined sections of p-type and n-type semiconductor

the ions *do not* leave their lattice positions and therefore do not constitute a current. The cores and the ions may be considered to be stationary. The holes and the excess electrons would be seen to move at random within the material. The movement of the holes and the electrons is due to thermal energy; this movement of charges in the absence of an applied field is called *diffusion.* (Diffusion is the movement of carriers from a region of high concentration to a region of lower concentration.) Even though the holes are in motion they are evenly distributed throughout the p-type semiconductor; the excess electrons are evenly distributed throughout the n-type semiconductor.

b. Figure 2–12 shows the same two sections of semiconductor (*a* above) joined to form a p-n junction.

1. Note that no external circuits or voltages have been connected to the material; nor is the material exposed to external electric or magnetic fields.

2. One would normally expect the holes in the *p*-type semiconductor and the electrons in the *n*-type semiconductors to flow toward each other, combine, and eliminate all holes and excess electrons. When the two types of semiconductor are joined, however, after a few combinations of holes and electrons result, a restraining force is set up automatically to preclude total combination. This restraining force is called a barrier. The cause and nature of the barrier are discussed in paragraph 2–15.

2–15. Junction Barrier

a. When the *p*-type semiconductor and the *n*-type semiconductor are joined (Fig. 2–12) some of the holes in the *p*-region and some of the excess electrons in the *n*-region diffuse toward each other and combine. Each combination eliminates a hole and an *excess* electron; the excess electron is now part of an electron-pair bond. This action occurs for a short time in the immediate vicinity of the junction. Negative acceptor ions in the *p*-region and positive donor ions in the *n*-region and near the junction are left uncompensated. Additional holes that would diffuse into the *n*-region are repelled by the uncompensated positive charge of the donor ions. Electrons that would diffuse into the *p*-region are repelled by the uncompensated negative charge on the acceptor ions. As a result, *total* recombination of holes and electrons cannot occur.

b. The region containing the uncompensated acceptor and donor ions is referred to as the *depletion region,* i.e., there is a depletion of holes and a depletion of excess electrons in this region. Since the acceptor and the donor ions are immobile (fixed) and are charged electrically, the depletion region is also referred to as the *space-charge region.* The electric field between the acceptor and the donor ions is called a *barrier.* The effect of the barrier is represented by the imaginary space-charge equivalent battery. The physical distance from one side of the barrier to the other is referred to as the *width* of the barrier, the width of the barrier depending on the density of holes and excess electrons in the germanium crystal. The difference of potential from one side of the barrier to the other is referred to as the *height* of the barrier. The height of the barrier is the intensity of the electric field (voltage of space-charge equivalent battery) and is measured in volts. With no *external* batteries connected, the barrier height is on the order of tenths of a volt.

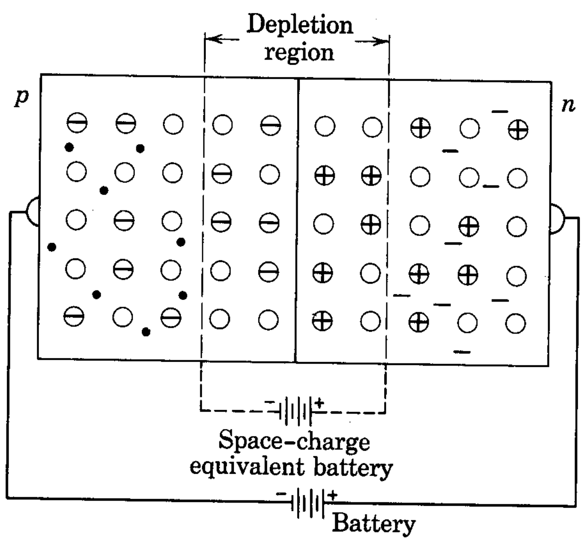
c. It is stated in *a* above that total recombination of electrons and holes cannot occur. Inspection of the polarity of the space-charge equivalent battery confirms this statement. Note that the electrons in the *n*-type germanium are already at the *highest* positive potential (positive terminal of

space-charge equivalent battery) with the crystal. The holes in the *p*-type germanium also are at the highest negative potential (negative terminal of space-charge equivalent battery) within the crystal. This condition precludes the movement of holes or electrons across the *p-n* junction.

2–16. *p-n* Junction, Reverse Bias

a. Figure 2–13 shows what happens when an external battery with the indicated polarity is connected to a *p-n* junction. Note that the negative terminal of the battery is connected to the *p*-type semiconductor and that the positive terminal of the battery is connected to the *n*-type semiconductor. The holes are attracted toward the negative terminal and away from the junction. The electrons are attracted toward the positive terminal and away from the junction. This action widens the depletion region and increases the barrier height (potential). Compare the width of the depletion region of Fig. 2–12 with that of Fig. 2–13.

b. Since the depletion region widens until the barrier height (potential of space-charge equivalent battery) equals the potential of the external



Legend:

○ Semiconductor core

⊖ Acceptor ion (negative)

⊕ Donor ion (positive)

• Hole (positive)

— Excess electron (negative)

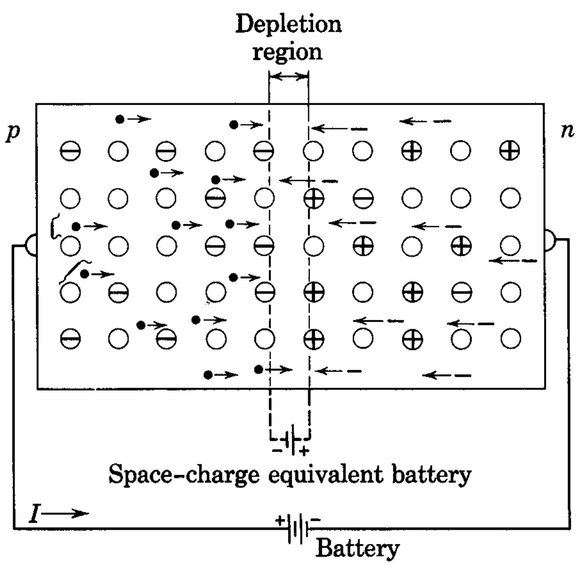
FIG. 2–13. *p-n* junction showing reverse bias

battery, no current flow of holes or electrons occurs because the battery voltages are in opposition. In this condition, the *p-n* junction is biased in the *reverse* direction; or simply, a *reverse bias* is placed across the *p-n* junction.

c. It is possible to apply a reverse bias greater than the largest possible barrier height. However, if this is done, the crystal structure will break down. In normal applications, this condition is avoided. The crystal structure will return to normal when the excess reverse bias is removed, provided that overheating does not permanently damage the crystal.

2–17. *p-n* Junction, Forward Bias

a. Figure 2–14 shows what happens when an external battery with the indicated polarity is connected to a *p-n* junction. Note that the positive



Legend:

○ Semiconductor core

⊖ Acceptor ion (negative)

⊕ Donor ion (positive)

• Hole (positive)

– Excess electron (negative)

Electron from electron-pair bond

FIG. 2–14. *p-n* junction showing forward bias

terminal of the battery is connected to the *p*-type semiconductor and the negative terminal of the battery is connected to the *n*-type semiconductor. The holes are repelled from the positive terminal of the battery and drift toward the junction. The electrons are repelled from the negative terminal of the battery and drift toward the junction. Because of their acquired energy, some of the holes and the excess electrons penetrate the depletion (space-charge) region and combine.

b. For each combination of an excess electron and a hole that occurs, an electron from the negative terminal of the external battery enters the *n*-type semiconductor and drifts toward the junction. Similarly, an electron from an electron-pair bond in the crystal, and near the positive terminal of the external battery, breaks its bond and enters the positive terminal of the external battery. For each electron that breaks its bond, a hole is created which drifts toward the junction. Recombination in and about the space-charge region continues as long as the external battery is applied.

c. Note that there is a continuous electron current (I) in the external circuit as indicated by the arrow. The current in the *p*-type semiconductor consists of holes; the current in the *n*-type semiconductor consists of electrons. In this condition, the *p-n* junction is said to be biased in the forward direction. If the forward bias is increased, the current increases.

Note: Throughout this text *current flow* refers to *electron current* rather than *conventional current.*

d. In paragraph 2–15*b* it was stated that the barrier potential with no external battery connected is on the order of tenths of a volt. It would appear, therefore, that an external battery of very low voltage (about 1 volt) would eliminate the barrier completely. However, the larger the voltage of the external battery, the greater the current flow through the crystal. Since the crystal has a relatively high resistivity, increased current causes increased voltage drop on both sides of the barrier. The remaining voltage of the external battery does not overcome the barrier completely. Normally, 1 to 1½ volts is used to bias the *p-n* junction in the forward direction. If excessive forward bias is used, excessive current will cause excessive thermal agitation and breakdown of the crystal structure.

2–18. Diode Action

a. Paragraphs 2–14 through 2–17 cover the mechanism of rectification through a *p-n* semiconductor diode. Figure 2–15 is a plot of current flow versus voltage applied to a practical *p-n* junction. Note that current flow in the forward bias direction is quite high (measured in ma). However, current flow in the reverse bias direction, although low (measured in μa), is not zero as might have been expected (par. 2–16).

1. In the *p*-type semiconductor some electrons in the electron-pair bonds gain enough energy to break and move out of the electron-pair bond structure. This action produces some additional holes which add to the existing holes caused by the acceptor impurities. The electrons that have broken their bonds become excess (free to move) electrons; these electrons are called *minority* carriers in the *p*-type semiconductor because they are always outnumbered by the holes which are referred to as *majority* carriers. The energy required to release an electron from an electron-pair bond may be in the form of heat, light, or other radiation. The opposite action can also occur; i.e., the excess electron can lose energy and fall into a hole, thus eliminating the excess electron and the hole. This action of generation and elimination of excess electrons and holes goes on continuously within the semiconductor material even at average room temperature.

2. In the *n*-type semiconductor the electrons that gain enough energy (1 above) to break their bonds, create excess electrons that add to the *majority* carriers (electrons provided by the donor impurities). The resultant holes, always outnumbered by the excess electrons, are referred to as the *minority* carriers.

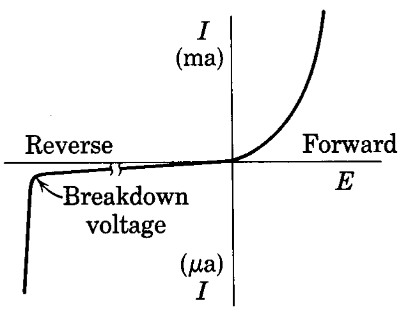


FIG. 2–15. Chart of current through, and the voltage across, a *p*-*n* junction

b. When the *p*-*n* junction is biased in the reverse direction for the majority carriers (par. 2–16), the *p*-*n* junction is biased in the *forward* direction (par. 2–17) for the minority carriers, electrons in the *p*-type semiconductor and holes in the *n*-type semiconductor. The internal mechanism of conduction for the minority carriers when *forward* biased (majority carriers reverse biased) is identical with that for forward-biased majority carriers.

c. Note that when a very high reverse bias is applied (Fig. 2–15), a high reverse current flows. This high current is not due to the minority carriers. A breakdown of the single crystal structure occurs (par. 2–16*c*).

SECTION IV. *p*-*n* JUNCTION, ENERGY LEVEL CONSIDERATIONS

2–19. General

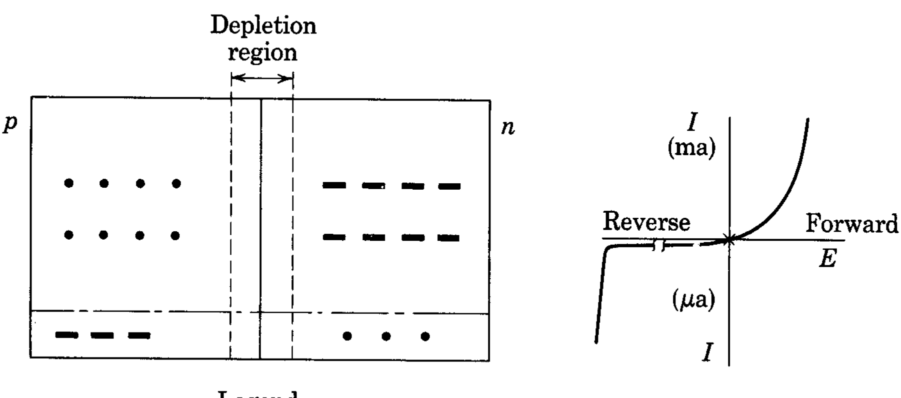
a. The *p*-*n* junction conduction and nonconduction mechanisms as discussed in paragraphs 2–13 through 2–18 are based purely on the attraction and repulsion of unlike and like charges, respectively. To that extent the

discussions are valid. However, to understand the difference in the conduction mechanisms between the normal diode and the tunnel diode, it is necessary to reconsider the normal diode conduction mechanism to take into account relative energy levels of holes and excess electrons. These energy levels during normal diode action are discussed in paragraphs 2–20, 2–21, and 2–22.

b. Figures 2–16, 2–17, and 2–18, showing no bias, reverse bias, and forward bias correspond, respectively, to Figs. 2–12, 2–13, and 2–14. Figures 2–12, 2–13, and 2–14 are more realistic in that they show more accurately the spacial distribution of holes and excess electrons throughout the semiconductor. Figures 2–16, 2–17, and 2–18 line up the holes and excess electrons according to their relative energy levels. This line-up of current carriers helps visualize the effect of energy levels on *p-n* junction action. For greater clarity these figures do not show semiconductor atoms or impurity ions. Additional simplification is obtained by separating the minority and majority carriers by a long-short dash line.

2–20. Energy Levels, Unbiased *p-n* Junction

Figure 2–16 shows essentially the same conditions as Fig. 2–12 when a region of *p*-type semiconductor and *n*-type semiconductor are joined. In the usual manner a depletion region (par. 2–15) is formed which acts as a barrier to the net flow of holes or excess electrons across the junction.



Legend:
- • Hole (positive)
- ▬ Excess electron (negative)

Majority carriers above long-short dash line
Minority carriers below long-short dash line

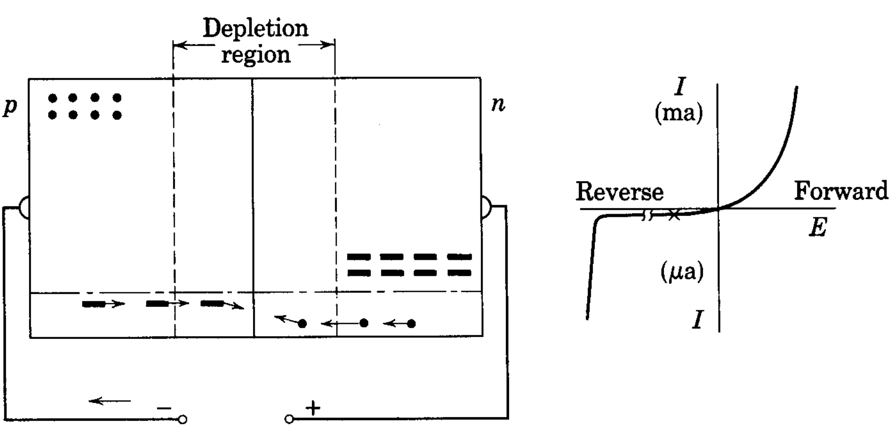
FIG. 2–16. *p-n* junction, unbiased, showing relative electron-energy levels of majority carriers and relative electron-energy levels of minority carriers

a. Assume that the long-short dash line is the lowest electron energy level for the majority carriers and that electron energy increases vertically up. Then the condition of equilibrium shown here is a small but equal number of majority holes and majority excess electrons at a medium energy level. The × on the current-voltage chart shows zero current flow caused by majority carriers.

b. The same condition (*a* above) prevails for the minority carriers. The lowest electron-energy level for the minority carriers is considered to be the bottom edge of the device. Note that relative energy levels between minority and majority carriers are not demonstrated in this diagram because they are not important to the discussion. As in the case of the majority carriers there is no net current flow across the barrier due to the minority carriers.

2–21. Energy Levels, Reverse-biased *p-n* Junction

Figure 2–17 shows what happens when a reverse bias (negative to *p*-type material and positive to *n*-type material) is applied. As previously explained (par. 2–16), the depletion width increases because the majority electrons move toward the positive terminal and the majority holes move toward the negative terminal. There is no current flow due to majority



Legend:

- • Hole (positive)
- ▬ Excess electron (negative)

Majority carriers above long-short dash line

Minority carriers below long-short dash line

Fig. 2–17. *p-n* junction, reverse biased, showing relative electron-energy levels of majority carriers, and relative electron-energy levels of minority carriers

carriers. There is a small current flow due to opposite movements of the minority carriers. This reverse-biased current is marked × on the current-voltage chart.

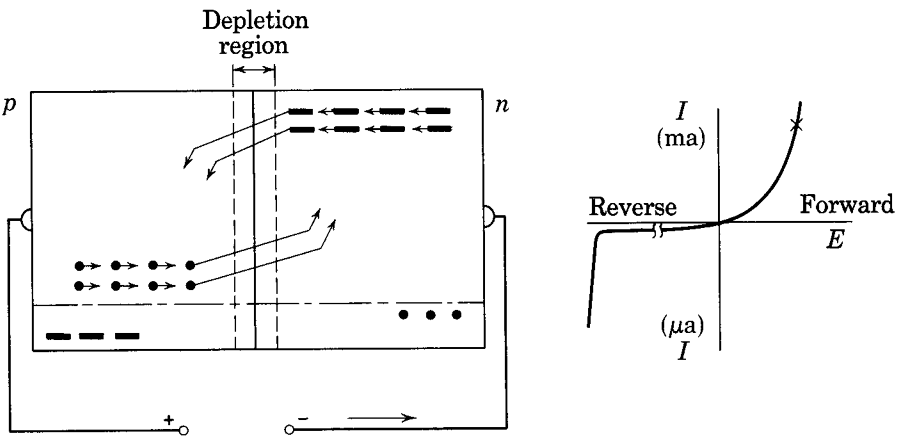
a. The same phenomenon is explained in terms of the energy levels of the carriers. If a battery were placed across an ordinary conductor, electrons would flow from the negative terminal of the battery, through the conductor, to the positive terminal of the battery. The reason for the current flow is that the *negative* potential excites the electrons to a high energy level proportional to the magnitude of the negative potential. With the greater energy, the loosely held electrons of the conductor atoms leave their atoms. Electrons will always move toward the lowest energy level; the positive terminal of the battery represents the lowest electron-energy level. This phenomenon is comparable to water seeking its lowest level. Water at a high level, if free to flow, will flow downhill. In its movement downhill, the water gives up energy. This energy is often used to drive hydroelectric plants.

b. Note what happens to the electron-energy levels of the majority carriers (Fig. 2–17) when a reverse bias is applied. The majority electrons, at the highest positive battery potential, are also at the *lowest* electron-energy level. *The electron-energy level of the hole indicates the energy level that electrons would assume if they occupied the position of the hole.* Since electrons seek the *lowest* energy level, the majority electrons do not move into the holes. No current flow due to majority carriers can be expected any more than water can be expected to flow uphill.

c. In the case of the minority carriers, however, the situation is reversed. The minority electrons are raised to a high energy level by the negative potential of the battery; the holes are reduced to a low electron-energy level by the positive terminal of the battery. The minority electrons therefore will readily flow from their high energy level to the low energy level of the holes. A small external current in the direction shown can be detected in the external circuit.

2–22. Energy Levels, Forward-biased *p-n* Junction

a. Figure 2–18 shows a forward-biased *p-n* junction (positive terminal of battery to *p*-type semiconductor, negative terminal to *n*-type semiconductor). Because of the negative potential of the battery, the majority electrons are raised to a high electron-energy level. The positive potential of the battery places the majority holes at a low electron-energy level. The majority electrons and the majority holes will readily flow toward the barrier. Within and about the barrier, numerous combinations of holes and electrons take place. This current flow due to a forward bias *high enough* (approximately 0.5 to 1 volt) to excite the majority electrons to an energy



Legend:

- • Hole (positive)
- ▬ Excess electron (negative)

Majority carriers above long-short dash line

Minority carriers below long-short dash line

FIG. 2–18. *p-n* junction, forward biased, showing relative energy levels of majority carriers and relative energy levels of minority carriers

level sufficient to overcome the barrier in large numbers is referred to as an *injection* current. A large external current flows in the direction indicated by the arrow. This current is marked by a $\times$ on the current-voltage chart.

b. The applied forward bias causes the minority electrons to assume a low energy level and the minority holes to assume a high energy level. Minority current flow therefore does not occur.

SECTION V. p_T-n JUNCTION, TUNNEL DIODES

2–23. General

a. The normal rectifying diode discussed in paragraphs 2–19 through 2–22 uses semiconductor materials lightly doped with one impurity atom for 10^7 (10 million) semiconductor atoms. If the semiconductor materials forming a junction are doped to the extent of *1000 impurity atoms* for 10^7 semiconductor atoms, the resultant current-voltage characteristic appears as shown in Fig. 2–19. Compare Figs. 2–15 and 2–19. The most important aspect of this characteristic (Fig. 2–19) is the peak current (I_p) rise with a small applied forward bias, the decreasing current with *increasing* forward

bias to a minimum (valley) current (I_v), and finally a normal increasing current with further increasing voltage. That portion of the characteristic between I_p and I_v represents an *ac negative resistance* which permits the use of this device in circuits requiring ac amplification (par. 3–6 through 3–34). The internal conduction mechanism that causes this phenomenon is discussed in paragraphs 2–24 through 2–28.

b. The most important effect of heavy doping of semiconductor materials is the effect of the quantity of doping on the width of the depletion region. The heavier the doping, the narrower the depletion region. Figure 2–20A shows a *p-n* junction formed by lightly doped semiconductor materials. For clarity, the holes, excess electrons, and electron-pair bonds are not shown. If heavier doping is used (Fig. 2–20B), the width of the depletion region is reduced. If the two types of semiconductor materials contain different amounts of impurities, the portion of the depletion region in the more heavily doped material is narrower than that in the lightly doped material (Fig. 2–20C). The reason for these conditions is that the barrier (electric field of the depletion region) is formed by a given number of donor ions on one side of the barrier and an *equal* number of acceptor ions on the other side of the barrier. The heavier the impurity concentration in the semiconductor material, the smaller the depth of penetration on that side of the junction necessary to establish a given number of uncompensated ions.

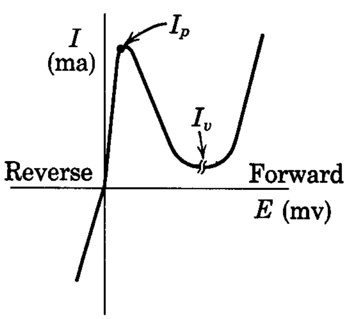


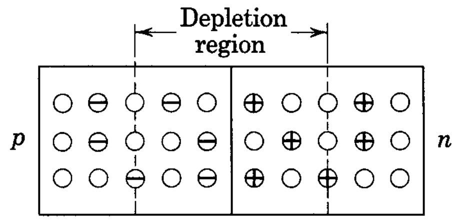
FIG. 2–19. Current-voltage chart of heavily doped (tunnel) diode

c. Because of the heavy doping used in tunnel diodes, the depletion region is only *one millionth of an inch* wide. This fact is important in explaining the current-voltage characteristic (Fig. 2–19) of the tunnel diode.

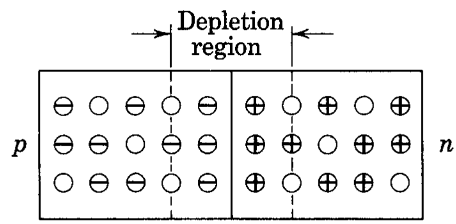
2–24. p_T-n Junction, Zero Bias

a. Figure 2–21 shows a junction formed by two heavily doped semiconductor materials. As in the case of the normal *p-n* junction (par. 2–20), a depletion region is formed even with zero bias. Note, however, that the depletion region is very narrow; compare Fig. 2–21 with Fig. 2–16. This ultra-thin depletion region is referred to throughout this text as a p_T-n junction.

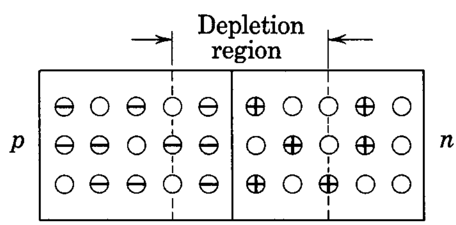
b. Note that a *large* number of majority holes and excess electrons are at the same electron-energy levels (par. 2–20). The same condition is true of the minority carriers. If there is any movement of carriers across the



A. p-and n-Type materials lightly doped



B. p-and n-Type materials more heavily doped



C. p-Type materials more heavily doped
n-Type materials lightly doped

Legend:

○ Semiconductor core

⊖ Acceptor ion (negative)

⊕ Donor ion (positive)

Note: Electrons and holes not shown

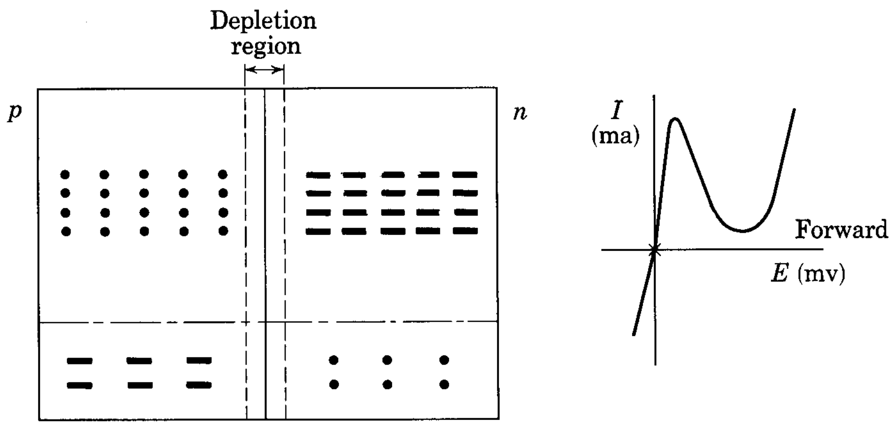
FIG. 2–20. *p-n* junctions, showing variation of depletion-region width with amount of doping of semiconductor materials

depletion region due to thermal energy, the net current flow will be zero because an equal number of like charges will flow in opposite directions. This zero net current flow is marked by an × on the current-voltage chart.

2–25. Junction with Forward Bias and Peak Current

a. Figure 2–22 shows a p_T-n junction with a small forward bias (about 50 mv) applied. The negative potential of the battery connected to the n-type material slightly raises the electron-energy levels of the majority

electrons. The positive potential of the battery connected to the p-type material slightly lowers the electron-energy level of the majority holes. The depletion region is also slightly narrowed. For normal p-n junctions no net current flow would occur since approximately 0.5 to 1 volt forward bias is required. However, a substantial current flow is measured externally in the direction of the arrow. The magnitude of current is marked by an $\times$ on the current-voltage chart.



Legend:

• Hole (positive)

▬ Excess electron (negative)

Majority carriers above long-short dash line

Minority carriers below long-short dash line

FIG. 2–21. p_T-n junction, showing narrow depletion region, energy level of carriers, and current-voltage chart

b. The phenomenon of high current with such small bias is referred to as *electron tunneling. Electrons will penetrate an ultra-thin barrier that normally they could not penetrate provided that there are vacancies (holes) on the opposite side of the barrier at the same electron-energy level as the penetrating electrons.* Note that the top row of majority electrons that move toward the barrier are deflected *back* into the n-type material because no vacancies (holes) of equal electron-energy level are present in the p-type material. The bottom three rows of majority electrons and the top three rows of majority holes do combine and cause current flow externally in the usual manner.

c. The majority electrons that penetrate the barrier appear in the p-type

material at the same energy level they had in the n-type material. This action is comparable to a man walking through a *tunnel* in a mountain that he could not possibly climb over. Thus the term *electron tunneling;* the electrons penetrate a barrier which normally they could not penetrate and appear at the same energy level on the other side of the barrier.

d. No current flow is caused by the minority carriers.

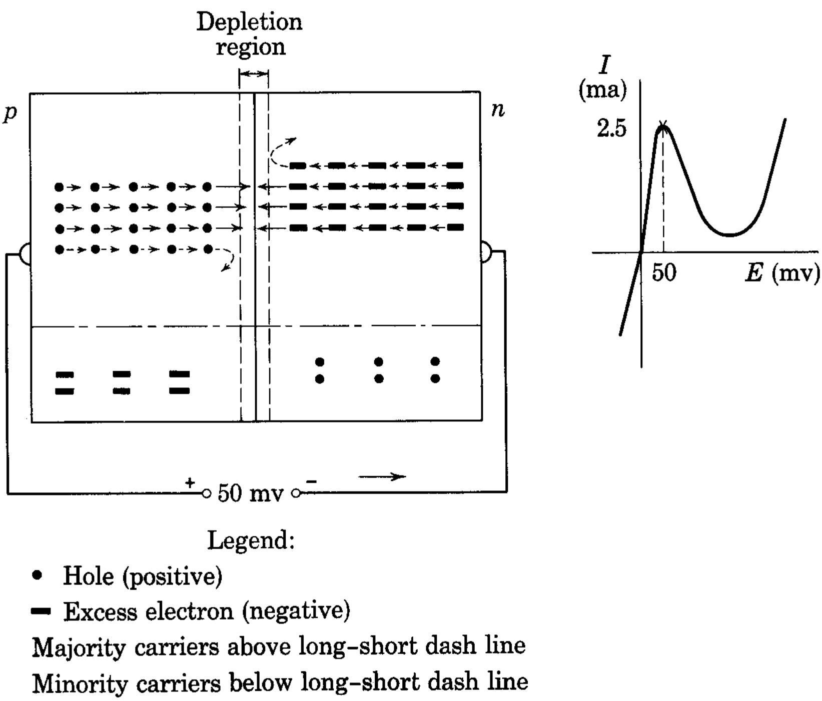
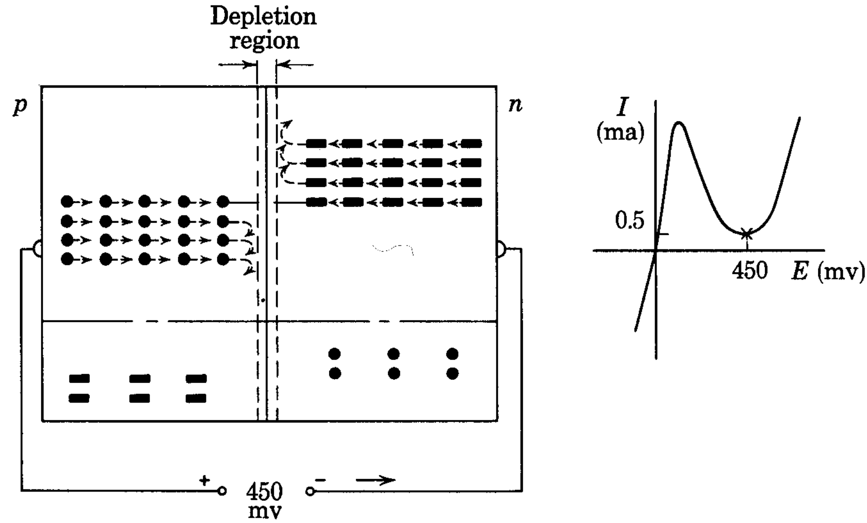


FIG. 2–22. p_T-n junction, slight forward bias, showing electron tunneling and peak current on current-voltage chart

2–26. p_T-n Junction with Forward Bias and Valley Current

a. Figure 2–23 shows a p_T-n junction with forward bias increased [from 50 mv (par. 2–25)] to 450 mv. The negative potential of the battery further increases the electron-energy level of the majority electrons; the positive potential further decreases the electron-energy level of the majority holes. Note that only one row of majority electrons and majority holes are at the same energy level. Very little electron tunneling (par. 2–25) occurs and only a small current (marked by × on the current-voltage chart) can be measured in the external circuit. This current is referred to as the



Legend:

- ● Hole (positive)
- ▬ Excess electron (negative)

Majority carriers above long-short dash line

Minority carriers below long-short dash line

FIG. 2–23. p_T-n junction, increased forward bias, showing little electron tunneling and valley current on current-voltage chart

valley current. Most of the majority electrons and holes are deflected back from the barrier.

b. No current flow is caused by the minority carriers.

2–27. p_T-n Junction with Forward Bias and Injection Current

a. Figure 2–24 shows the p_T-n junction with forward bias increased [from 450 mv (par. 2–26)] to 500 mv. The negative potential of the battery further raises the electron-energy level of the majority electrons; the positive potential further decreases the electron-energy level of the holes. All of the majority electrons have gained sufficient energy to penetrate the barrier and combine with the holes of low electron-energy level. The resultant current flow is the same as that which occurs in a normal p-n junction (par. 2–22) and is referred to as *injection* current to distinguish it from *tunnel* current.

b. The forward-bias voltage at which the injection current occurs in a p_T-n junction is approximately the same voltage at which injection current

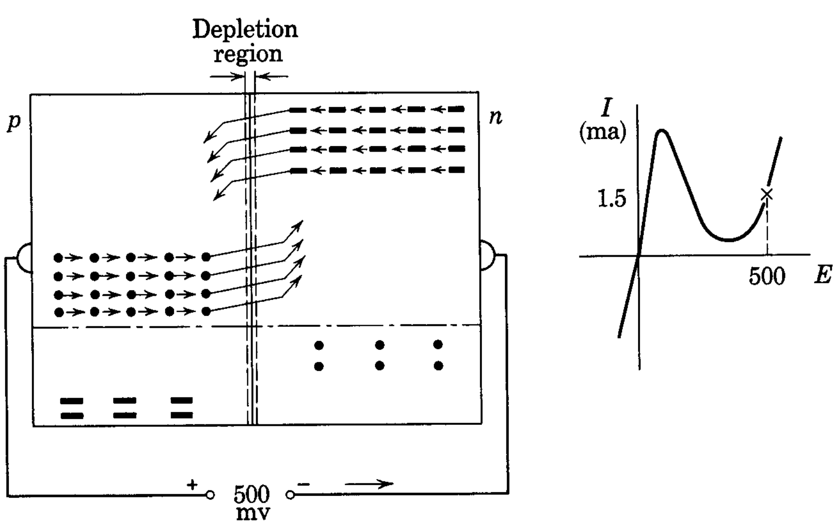
would occur in a *p-n* junction made of the same semiconductor material.

c. The current flow from zero bias through the peak current down to the valley current is called tunnel current. Beyond the valley current, it is called injection current.

2-28. p_T-n Junction, Reverse Biased

a. Figure 2-25 shows a p_T-n junction reverse biased. The majority electrons are greatly reduced in electron-energy level and the majority holes are greatly increased in electron-energy level. No current flow due to majority carriers occurs.

b. The application of negative potential to the *p*-type material raises the electron-energy level of the minority electrons originally present. It also increases the number of minority electrons by causing electrons from electron-pair bonds to break their bonds and become excess electrons. An equal number of holes are also formed, and these join the original number of majority holes. Compare the numbers of majority holes and minority electrons in Figs. 2-24 and 2-25. This generation of holes and excess elec-



Legend:

- Hole (positive)
- Excess electron (negative)

Majority carriers above long-short dash line

Minority carriers below long-short dash line

FIG. 2-24. p_T-n junction, forward biased to cause injection current as marked on current-voltage chart

trons occurs only because the semiconductor is heavily doped and acts very much like an ordinary conductor.

c. Although the width of the depletion region has been slightly increased by the reverse bias, the increase is not sufficient to prevent electron tunneling between the minority electrons and the minority holes that are at

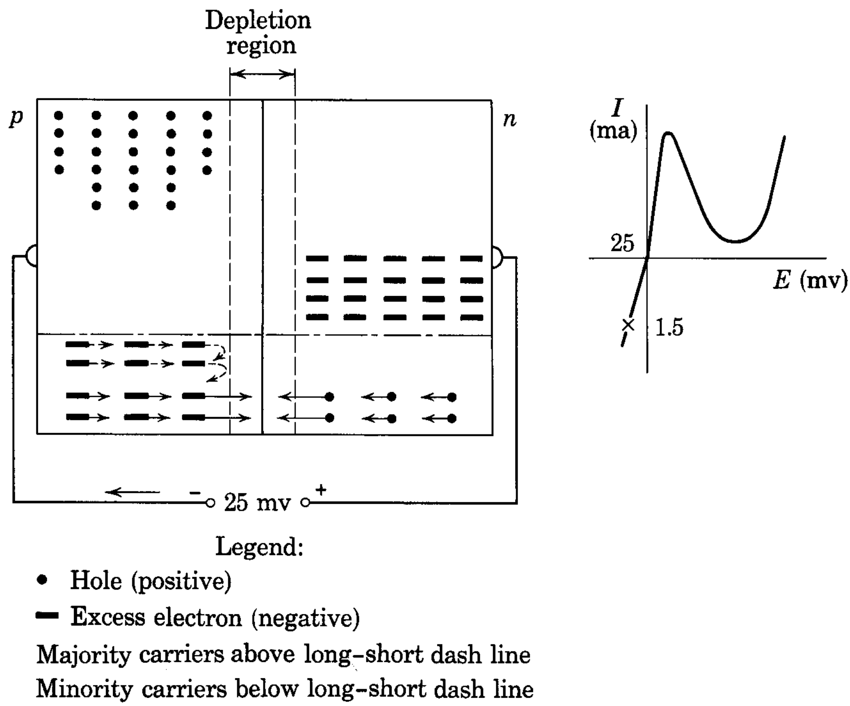


FIG. 2–25. p_T-n junction, reversed biased, showing resultant current marked by × on current-voltage chart

the same electron-energy level. As a result, a relatively heavy reverse-bias current (marked by an × on the current-voltage chart) flows in the direction of the arrow with only 25 mv applied.

2–29. Peak and Valley Currents

a. The peak and valley currents of p_T-n junctions made from a particular semiconductor material always occur at the same peak voltage and valley voltage, respectively. Figure 2–26A shows the forward current-voltage chart of two tunnel diodes, each made of germanium. The differences in magnitude in peak currents shown for curve 1 and curve 2 can be achieved by

increasing or decreasing the amount of doping, respectively, or by varying the size of the junction area. Units varying in peak currents from 10 μa to 10 amp have been constructed. However, most present applications use units having peak currents of from 1 to 50 ma. The ratio of peak current to valley current ordinarily remains the same. The valley current is usually 10% of the peak current. Compare the magnitudes of peak to valley currents for each curve shown. The voltage at which the injection current equals the peak current is referred to as the injection-current (or forward) voltage. For germanium units this voltage occurs approximately between 500 mv and 600 mv.

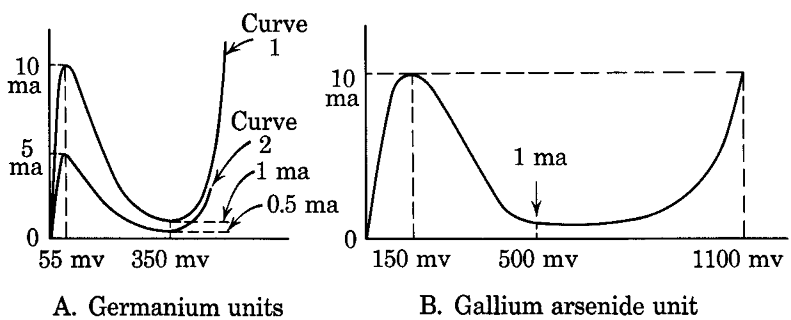


FIG. 2–26. Voltage-current charts for germanium and gallium arsenide tunnel diodes

b. For gallium arsenide unit (Fig. 2–26B), the peak current occurs at 150 mv, the valley current at 500 mv, and the injection current voltage at 1100 mv. This spread of approximately one volt between peak-tunnel voltage and injection-current voltage promises a wider use for gallium arsenide units, particularly in computer switching applications.

c. The fact that the peak-tunnel current and the valley current occur at the same respective voltages for a tunnel diode made of a particular material might have been expected from the theory of electron tunneling and injection (pars. 2–24 through 2–28). These conduction mechanisms depend upon specific energy levels of current carriers, which, in turn, are affected by the applied voltage and the forces existing between the orbiting electron and its nucleus.

SECTION VI. TUNNELTRONS

2–30. Tunneling Through Ultra-Thin Insulators

a. In experiments conducted at the General Electric Company, a scientist, Ivar Giaever, discovered that the phenomenon of tunneling occurs in

devices other than semiconductors. In these experiments films of such conductors as aluminum, lead, and tin were separated by ultra-thin insulators—aluminum oxide, tantalum oxide, and nickel oxide (Fig. 2–27A). Two conductors separated by an insulator normally form a capacitor even if the insulator is only one ten-thousandth of an inch thick. In these experiments the insulating film was only 10 to 100 atoms thick. Furthermore, the tunnel effect (using the materials cited above) occurred only if the

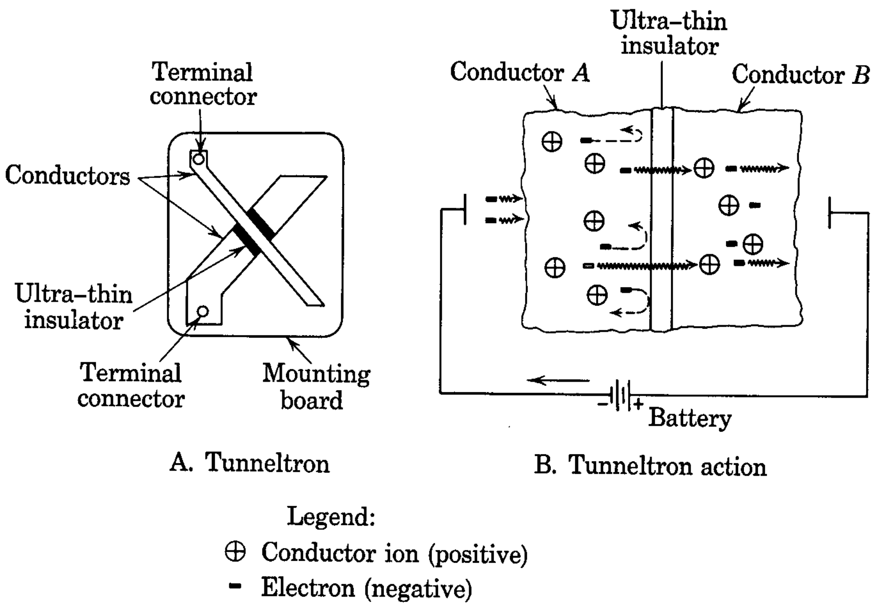


FIG. 2–27. Physical construction of tunneltron and tunneltron action

conductors were made superconducting (negligible in resistance) by cooling to liquid helium temperatures. The resultant device has been referred to as a *tunneltron*. (*Note:* Subsequent announcements by scientists at the Republic Aviation Corporation indicate the development of tunneltrons that operate and display the tunneling phenomenon at room temperatures. These devices use films of titanium as the conductors and a film of glass as the insulator.)

b. A diagrammatic representation of the theory of operation of the tunneltron is shown in Fig. 2–27B. All tunneling action takes place at the point where the three films of material overlap. A battery is connected across the tunneltron at the terminal connectors. For discussion purposes the conductor connected to the negative terminal of the battery is arbitrarily

designated conductor *A*; the other conductor is designated conductor *B*. Assuming in this case that the conducting material used is aluminum, then the positive conductor ion represents the nucleus of the aluminum atom and twelve of its thirteen orbiting electrons. When one of the three (valence) outer orbit electrons (Fig. 2–2B), moves away from the atom, the remainder of the atom (now a positive ion) exerts a strong holding influence on the other two valence electrons.

c. With battery voltage applied as shown (Fig. 2–27B) one electron from each aluminum atom in conductor *A* moves away from the negative terminal of the battery and toward the ultra-thin insulating barrier. The electrons in conductor *B* have a tendency to move toward the positive terminal of the battery. The majority of the electrons in conductor *A* that reach the insulating barrier are *reflected* back into conductor *A*. Some of the electrons penetrate the barrier and enter conductor *B*. Although the majority of the electrons are reflected back, the electrons are so numerous that an appreciable current flows. For each electron that penetrates the barrier, one electron enters conductor *A* from the negative terminal of the battery; one electron leaves conductor *B* and enters the positive terminal of the battery.

d. At very low voltages, current increases with applied voltage until a current peak is reached. Further increase in voltage results in lower current values until a current minimum occurs, after which current increases with applied voltage. The net result of this action gives rise to an ac negative resistance identical with that obtained with the tunnel diode.

2–31. Advantages and Disadvantages of Tunneltron

a. The major disadvantage of the tunneltron developed by Mr. Giaever is that one or both of the conductors must be in a state of superconductivity. Superconductivity is achieved by reducing the temperature of the conductor to near liquid helium temperatures. This requirement can be met by operating the tunneltron in a liquid helium refrigerator which in many electronic applications would not be practical economically. However, this major disadvantage is overcome if the tunneltron is made of titanium and glass.

b. One important property of superconducting metals is the ability to make them nonsuperconducting by the application of strong external magnetic fields. Because the ac negative resistance displayed by the tunneltron depends upon superconductivity of metals, it is obvious that variation of the ac negative resistance can be achieved by using magnetic fields. Such variation of ac negative resistance is not possible with the tunnel

diode. However, specific circuits employing this important characteristic of the tunneltron have not yet been devised.

c. Manufacturing processes for depositing films of metals in sandwiches of conductor-insulator-conductor as required by tunneltrons are standard and extremely inexpensive. It is expected therefore that costwise, tunneltrons can compete well with tunnel diodes.

d. At the present time the tunneltron is undergoing extensive study in an effort to determine specific and practical applications. At this writing no practical and commercial circuit applications of the tunneltron have been published. It is expected, however, that its properties will permit its application to low-noise amplifiers, high-frequency oscillators, and switching and memory circuits. Even if the tunneltron never emerges from the laboratory, its discovery has added greatly to the knowledge of tunneling phenomenon. If it does take its place among other practical active circuit devices, the discussions of ac negative resistance as presented in subsequent chapters of this text will be equally applicable to tunneltrons as well as tunnel diodes. Furthermore, it is expected that the tunneltron can, with minor circuit changes, substitute for the tunnel diode in many applications.

2-32. Comparison of Tunneltron and Tunnel Diode Actions

a. The basic tunneling concepts in the tunneltron and in the tunnel diode are identical; i.e., that electrons will move through an ultra-thin barrier provided there are vacancies (positive areas) on the other side of the barrier at the same energy level as the electrons. In the tunnel diode, the vacancy is a hole, a positive area in an electron-pair bond that appears to move freely. In the tunneltron the vacancy is a positive ion, an atom that has lost an electron and is not considered to be mobile. The hole is the result of the loss of an electron by a semiconductor (germanium, silicon, etc.) atom to an acceptor atom. The hole, therefore, although mobile, actually represents at any given instant of time a positive semiconductor ion. The positive area vacancies in the tunneltron and the tunnel diode then represent positive ions. The term *hole,* however, is reserved to designate a vacancy in the electron-pair bond structure of a crystal lattice network and is not applied to all positive ions.

b. The tunnel diode depends for its ultra-thin barrier on the depletion region between a heavily doped *p*-type semiconductor and a heavily doped *n*-type semiconductor; the barrier is created by an array of uncompensated negative acceptor ions on one side of a junction and uncompensated positive donor ions on the other side of the junction. The tunneltron depends for its ultra-thin barrier on a film of insulating material only several atoms thick.

2-33. Summary

a. Atoms are composed of positively charged particles called protons, negatively charged particles called electrons, and uncharged particles called neutrons.

b. A conductor is a material that has many loosely held electrons. Examples are silver, copper, and aluminum.

c. An insulator is a material that has few loosely held electrons. Examples are rubber, glass, and porcelain.

d. A semiconductor is a material, the resistivity of which is between those of conductors and insulators. Examples are germanium, silicon, and gallium arsenide.

e. A crystal is a material with atoms arranged in a specific pattern.

f. The properties of *polycrystalline* materials, e.g., copper and silver, are quite different from those of *single* crystal materials; single crystal materials are prepared for use in electron semiconductor devices. Germanium and silicon may be processed as single crystal materials.

g. Electrons shared by adjacent atoms in a crystal form electron-pair bonds.

h. *n*-Type germanium contains donor impurities. Donor impurities are materials that have *five* valence electrons, one of which cannot form an electron-pair bond. This electron is called an excess electron.

i. Arsenic, antimony, and phosphorus are examples of donor materials.

j. *p*-Type germanium contains acceptor impurities. Acceptor impurities are materials that have three valence electrons. Because *four* valence electrons are required to form and complete all adjacent electron-pair bonds, a hole is created.

k. Aluminum, gallium, boron, and indium are examples of acceptor impurities.

l. A hole can be considered a positive charge which diffuses or drifts through a crystal. The drift of holes constitutes a current.

m. A depletion (space charge) region occurs at a *p-n* junction. The potential difference across the depletion region is called a barrier. The width of the barrier is the width of the depletion region. The potential difference is called the height of the barrier.

n. Forward bias of a *p-n* junction causes heavy current (flow of majority carriers). Reverse bias causes very low current (flow of minority carriers).

o. p_T-n Junctions are formed by heavily doped *p*- and *n*-type semiconductor materials which result in ultra-thin (one millionth of an inch) depletion regions.

p. Electron tunneling will occur only through ultra-thin barriers provided that excess electrons on one side of the barrier and holes (vacancies) on the other side are at the same electron-energy level.

q. The most important property of the tunnel diode current-voltage chart is the region between the peak current (I_p) and the valley current (I_v), wherein current *decreases* with *increasing* voltage.

r. Decreasing current with increasing voltage represents an *ac negative resistance,* usable in electronic circuits to achieve gain.

s. The tunneltron, a *sandwich* of two conductors and an ultra-thin insulator, also displays the phenomenon of electron tunneling.

Chapter 3

AMPLIFICATION AND OSCILLATION USING AC NEGATIVE RESISTANCE

SECTION I. HISTORY AND DEVICES

3–1. General

Devices and circuit arrangements (other than tunnel diodes) displaying ac negative resistance and employed for amplification and oscillation were studied and applied as early as 1918. (See References, Appendix A.) One of the most popularly known devices is the thyratron, a gas-filled triode, which displays an N-shaped voltage-current characteristic. A circuit arrangement using an ordinary pentode electron tube (par. 3–2) and used as an oscillator also displays negative resistance; this circuit is called a *transitron*. A number of solid-state devices also display negative-resistance characteristics; these devices are discussed briefly in paragraph 3–3. None of these devices have found extensive use in electronics or offer promise of extensive use because not one has a combination of all the advantages of tunnel diodes (par. 1–4); namely, high-frequency oscillation, high-speed switching, simplicity of construction, and ease of manufacturing.

3–2. Transitron Oscillator

a. Figure 3–1A is a schematic diagram of a transitron oscillator. In this circuit, resistor $R1$ develops the ac suppressor voltage coupled from the screen grid through capacitor $C2$. Capacitor $C2$ is a blocking capacitor having negligible reactance at the oscillating frequency. Capacitor $C1$ and resistor $R2$ develop cathode bias. Coil $L1$ and capacitor $C3$ form a tank circuit and primarily determine the frequency of oscillation. Capacitor $C4$ couples the signal to the following stage.

b. The most important aspects of this circuit are:

1. The control grid is connected directly to the cathode. The cathode current, therefore, is a fixed quantity.

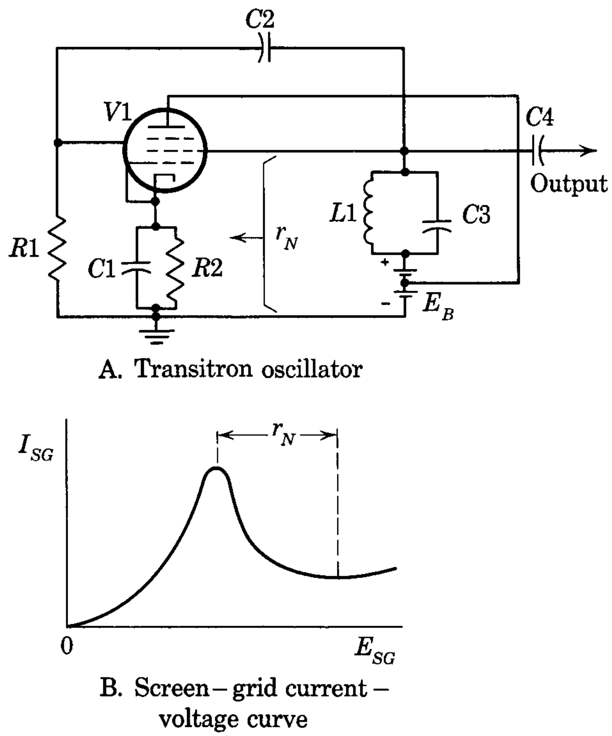


FIG. 3–1. Transitron oscillator and its screen current–screen voltage chart

2. The suppressor grid, normally at ac ground potential in amplifier circuits, is made to vary in potential with the voltage of the tank circuit.
3. The tank circuit ($L1$ and $C3$) is connected from *screen* to ac ground through the battery.
4. The plate is at a *lower* dc potential than the screen, and is at ac ground potential.

c. Figure 3–1B shows the screen current (i_{SG}) screen voltage (e_{SG}) chart. Note that for a portion of the curve (marked r_N) the screen current *decreases* with increasing screen voltage. This portion represents ac negative resistance which can sustain oscillation. When biased to operate on this portion of the curve, the tank circuit *sees* a negative resistance from screen grid to ac ground. If the total positive resistance (suppressor resistor $R1$, tank coil $L1$ resistance, and load resistance of following stage) in parallel with the tank circuit is greater in value than r_N, oscillations will be sustained. The latter statement will be more fully appreciated after reading paragraphs 3–26 and 3–27.

d. The negative resistance in the screen-grid circuit is caused by varying the voltage of the suppressor grid and placing the plate at a *lower* potential than the screen. The suppressor-grid voltage *does not* determine the number of electrons between the plate and the screen, but it does determine the *percentages* of electrons that arrive at the plate and the screen. When the suppressor grid goes more positive, a larger percentage of the electrons are urged on to the plate. When the suppressor grid goes more negative, a larger percentage of the electrons are deflected to the screen. The suppressor and the screen are at the same ac potential. Thus, any increase in suppressor-grid voltage increases screen voltage and decreases screen current. Any decrease in suppressor-grid voltage decreases screen voltage and increases screen current.

3–3. Semiconductor Devices Displaying Negative Resistance

In addition to tunnel diodes, the solid-state devices (Fig. 3–2) discussed below also display negative resistance. None of the following devices depend upon electron tunneling for its negative-resistance characteristic. As a result, all of them have the advantage of operating at voltages above those normal for tunnel diodes. However, each depends upon the lifetime of current carriers (holes and electrons) within the semiconductor material for its speed of operation. The lifetime of current carriers refers to the amount of time required before holes and electrons combine within the material. Under certain conditions (such as light doping) the lifetime of current carriers is relatively long (microseconds or tenths of a microsecond) and precludes use of these devices for the high-speed switching and high-frequency applications.

a. The point-contact transistor and its current-voltage chart display negative resistance as indicated (Fig. 3–2A). Its negative resistance depends upon current feedback from the output circuit (collector base) to the input circuit (emitter base).

b. The *p-n-p-n* transistor (Fig. 3–2B) has a current-voltage chart identical in shape with that of the point-contact transistor. It is believed that the bulk material of the point-contact transistor also has alternate layers of *p*-type and *n*-type materials. In each case current feedback from output to input circuits causes the negative resistance.

c. The four-layer diode (Fig. 3–2C) displays a negative resistance as indicated. A brief explanation of the operation of the four-layer diode and a sample application are covered in paragraph 3–4.

d. The unijunction transistor (Fig. 3–2D) exhibits a negative-resistance characteristic as indicated. The unijunction transistor is also referred to as a double-base diode. A brief discussion of its operation and a sample application are covered in paragraph 3–5.

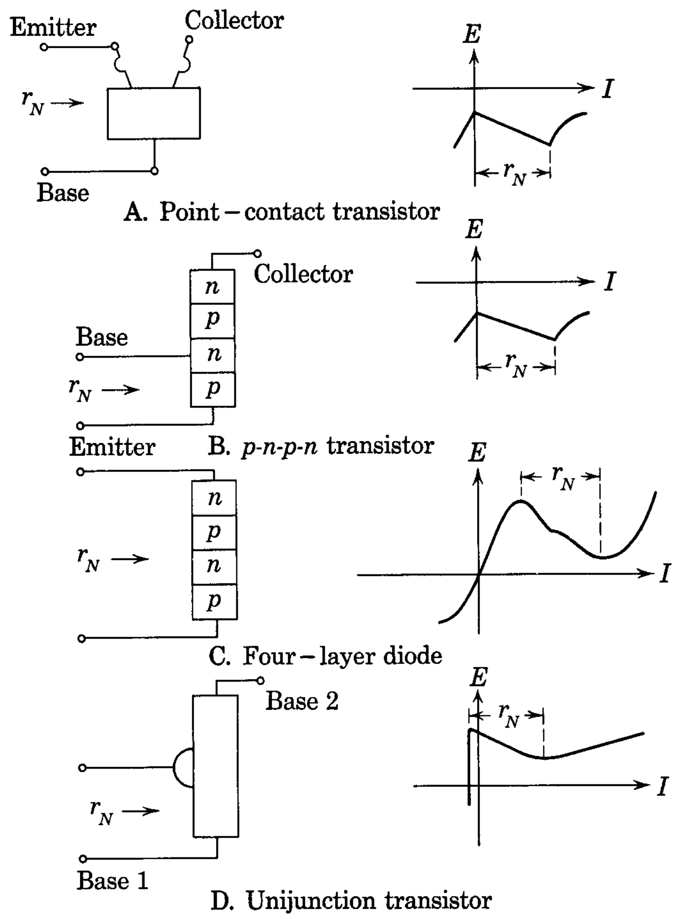


FIG. 3–2. Semiconductor devices and corresponding current-voltage charts, displaying negative resistance

3–4. Four-Layer Diode

a. Operation. The four-layer diode (Fig. 3–3A) consists of four layers of semiconductor material. The four layers of n-type and p-type material form three p-n junctions. When properly biased, the center p-n junction is reverse biased and the outer p-n junctions are forward biased. The emitter-base junction of a three-terminal transistor is always forward biased, and the collector-base junction is always reverse biased. A brief review of transistor fundamentals is covered in Appendix C. The four-layer diode, therefore, can be analyzed as two separate junction transistors. The dotted line through the n- and p-type materials in the center divides the four-layer

diode into a p-n-p transistor and an n-p-n transistor. The p-n junction of the p-n-p transistor (Fig. 3–3B) connected to the positive side of battery V_{EE} is forward biased and is therefore the emitter-base junction. The remaining p-n junction of the p-n-p transistor is reverse biased and is therefore the collector-base junction. The p-n junction of the n-p-n transistor connected to the negative side of battery V_{EE} is forward biased and is therefore the emitter-base junction. The remaining p-n junction of the n-p-n transistor is reverse biased and is therefore the collector-base junction. The base of the p-n-p transistor is connected directly to the collector of the n-p-n transistor. The collector of the p-n-p transistor is connected directly to the base of the n-p-n transistor. The schematic representation of the four-layer diode is shown in Fig. 3–3C. Proper biasing is obtained from battery V_{EE}. The arrows represent electron-current flow.

b. Application. The schematic diagram of a sawtooth oscillator employing the four-layer diode is shown in Fig. 3–3D. Battery V_{EE} furnishes power for the oscillator circuit. Switch $S1$ completes the circuit and applies power to the oscillator circuit. The time constant of resistor $R1$ and capacitor $C1$ determines the frequency of oscillation. The four-layer diode may

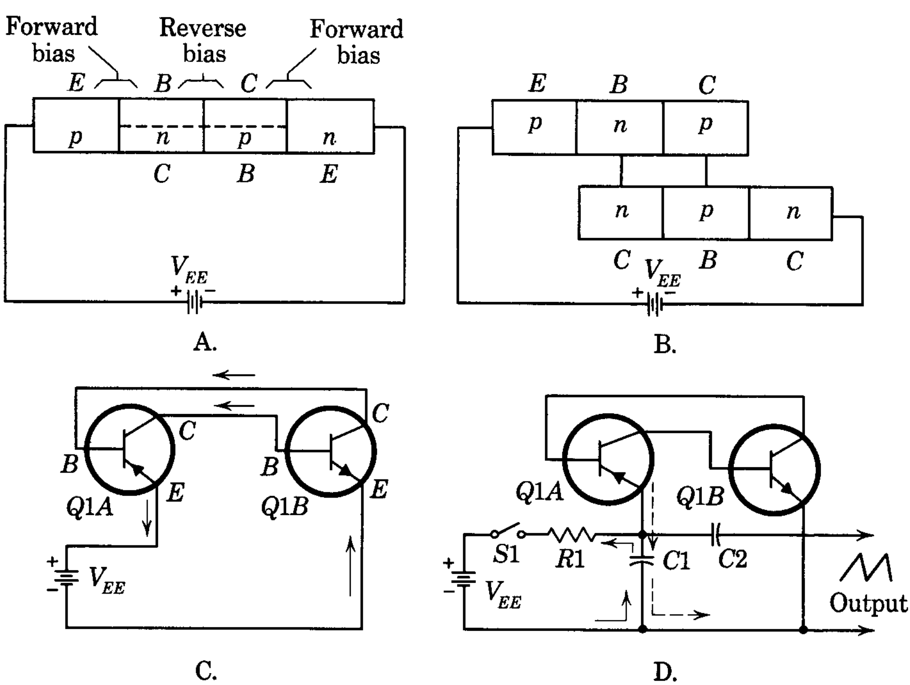


FIG. 3–3. Four-layer diode, and sawtooth generator

be considered the equivalent of two transistors ($Q1A$ and $Q1B$) so connected that the output of one feeds the input of the other. At a certain critical voltage (peak voltage, Fig. 3–2C), the current feedback from transistor $Q1B$ to transistor $Q1A$ becomes large. Both transistors conduct heavily and represent a very low dc resistance (high current, low voltage). When power is applied to the circuit through switch $S1$ (Fig. 3–3D), capacitor $C1$ appears as a short circuit. Electron current flows through resistor $R1$ and capacitor $C1$ charges (solid-line arrows). When the voltage across capacitor $C1$ reaches the critical voltage, heavy current flows through the four-layer diode. Capacitor $C1$ then discharges (dashed-line arrow) rapidly through the low resistance of the four-layer diode. As capacitor $C1$ discharges, the bias applied to the four-layer diode decreases. When the bias reaches a minimum value for conduction, the four-layer diode represents a very high resistance. Capacitor $C1$ again charges to a point where sufficient bias causes the four-layer diode to conduct heavily. This sequence keeps repeating, and the output, coupled through capacitor $C2$, is a sawtooth waveform.

3–5. Unijunction Transistor

a. Operation. The unijunction (single junction) transistor (Fig. 3–4A) is actually a diode with two connections made to one portion of the semiconductor. The unijunction transistor is also referred to as a double-base diode. If terminals 2 and 3 are connected, the resultant device would have the characteristics of a rectifying junction diode.

1. With battery V_{BB} connected as indicated, electron current flows from terminal 3 to terminal 2 as indicated. The usual resistance between these two terminals is 5 to 50 kilohms, depending on the particular unit. With terminal 1 connected to terminal 3, the p-n junction is reverse biased. The only current flowing through terminal 1 is a reverse-biased current; this is a negligible current consisting of electron flow (solid-line arrow) from terminal 1 to the p-n junction and holes (dashed-line arrow) from terminal 2 to the p-n junction. The junction represents a very high resistance.

2. Battery V_{BB} establishes within the n-type semiconductor the voltage gradient indicated from terminal 2 to terminal 3 (Fig. 3–4B). Battery V_{EE} is inserted in the circuit between terminals 1 and 3 with the polarity indicated. If the voltage of battery V_{EE} is less than the voltage gradient opposite the p-type material, the condition shown in Fig. 3–4A prevails. If the voltage of battery V_{EE} is greater than the voltage gradient opposite the p-type material, the p-n junction becomes forward biased. Heavy electron-current flow (solid-line arrows) occurs in the n-type material (Fig. 3–4B), and a heavy hole-current flow (dashed-line arrows) occurs

in the p-type material; electrons flow out of terminal 1. The junction represents a very low resistance.

3. These characteristics (1 and 2 above) of the unijunction transistor make it especially suitable for use in multivibrators and sawtooth generators (b below).

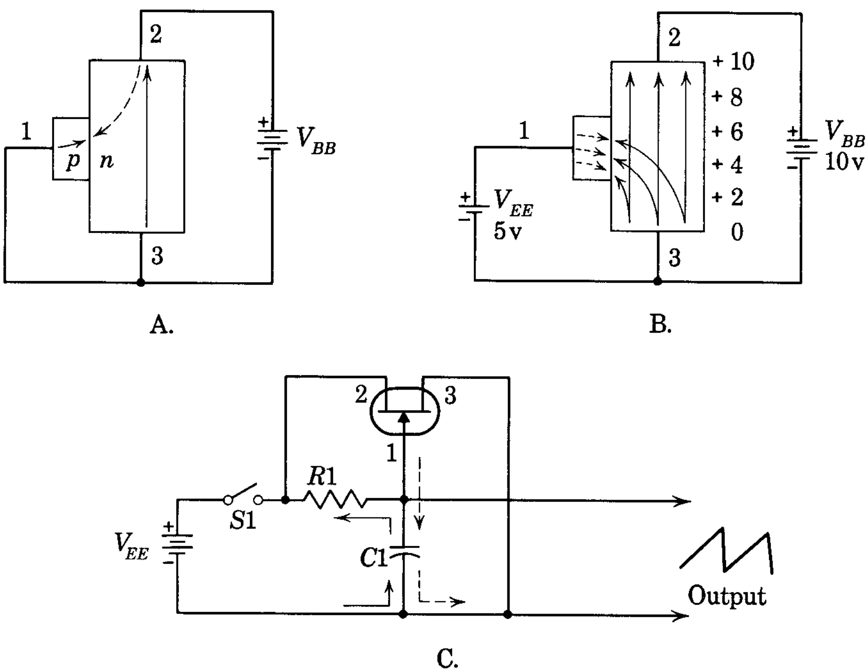


FIG. 3-4. Unijunction transistor and sawtooth generator

b. Application. A sawtooth generator using a unijunction transistor is shown in Fig. 3-4C. When power is applied to the circuit by closing switch $S1$, the conditions are the same as those described in $a1$ above. Capacitor $C1$ charges (solid-line arrows) very slowly through resistor $R1$. The charging time equals the product of the capacitance of capacitor $C1$ and the resistance of resistor $R1$. As capacitor $C1$ charges, the positive dc voltage across it rises. When this voltage is greater than the voltage gradient opposite the p-type material, the p-n junction is forward biased ($a2$ above), and the capacitor discharges (dashed-line arrows) very rapidly. The discharge time equals the product of the capacitance of capacitor $C1$ and the resistance of the forward-biased p-n junction. After capacitor $C1$

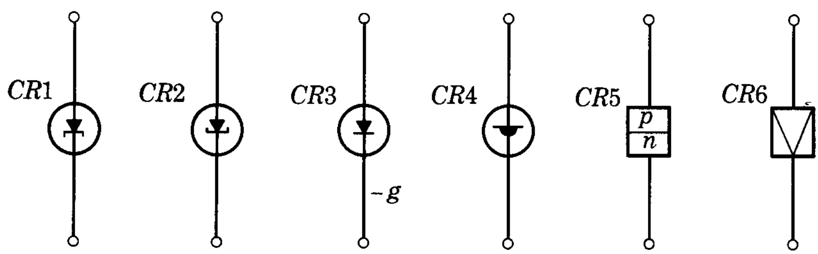
is discharged, the conditions are again the same as those described in *a*1 above. Capacitor *C*1 again charges and establishes a forward bias for the *p-n* junction. This sequence continues and the slow charging and rapid discharging of capacitor *C*1 produces a sawtooth waveform in the output.

SECTION II. AMPLIFICATION, SERIES ARRANGEMENT

3–6. Reference Designations and Graphical Symbols

a. Reference Designation. The reference designation for an electrical device consists of one or more letters and any whole number. In the absence

Anode terminals



Cathode terminals

FIG. 3–5. Tunnel diode graphical symbols

of an established standard, this text will continue the convention of using *CR* for any crystal diode; this includes rectifying diodes, and Zener diodes, as well as oscillating (frequency determining) crystals. Thus, throughout this text, tunnel diodes will be designated *CR*1, *CR*2, *CR*3, etc.

b. Graphical Symbol. Current technical literature includes a variety of graphical symbols for the tunnel diode. Several of these are illustrated in Fig. 3–5. No standard has been established for the tunnel diode at the time of this writing.

1. The illustrations in this text will employ the graphical symbol designated *CR*1. This symbol is that of a rectifying diode with the cathode portion formed into a "T." The direction of electron-current flow, when the device is forward biased, is opposite to that of the arrowhead.
2. The symbol designated *CR*2 is similar to *CR*1 except that the ends of the "T" are turned toward the arrowhead.
3. The symbol designated *CR*3 is that of a conventional diode with the letter $-g$ external to the symbol to indicate the negative conductance of the device.

4. The symbol designated *CR*4 represents a blunt arrowhead that has passed (or tunneled) through the junction to appear on the other side. This symbol might very well lend itself to future hybrid arrangements of transistors in which one or both junctions may be p_T-n junctions.

5. The symbols designated *CR*5 and *CR*6 are simple rectangular arrangements to represent the tunnel diode.

c. Conclusion. The symbol designated *CR*1 is being used in this text because it has been used often enough to indicate some preference for it. Each symbol has merit of its own, and the variety is a tribute to man's imagination. No effort will be made here to predict which symbol will be standardized eventually by the Institute of Radio Engineers; it might very well be one not illustrated here.

3–7. AC Negative Resistance

a. The concept of *ac negative resistance* is quite often difficult to grasp. Most of the difficulty arises from the fact that *negative resistance* is a *second-order* idea and not a *first-order idea or occurrence.* Here the term *second-order idea* simply means a *mathematical* method or technique used to represent this natural phenomenon quantitatively in equations and formulas. Such representation in turn permits the design engineer to predict the behavior of a device exhibiting "negative resistance" if the device is used in a given circuit; this procedure precludes the necessity of building the circuit and actually testing it in a laboratory. In other words, the mathematical representation eliminates to a great extent the need for trial-and-error procedures which are extremely costly.

b. Explanations of ac negative resistance in terms of this second-order (mathematical) idea gives rise to statements such as:

1. Negative resistance acts like a *source of power.*
2. Negative resistance acts like a *generator.*
3. Negative resistance, unlike positive resistance, which absorbs energy, *gives up energy.*

c. These statements are all true on a mathematical basis, but they do not enlighten the reader as to the first-order occurrence. The reader cannot readily reconcile himself to these statements because fundamentally he *knows* that the device that exhibits negative resistance *is not* a power source or a generator, and it does not *give up* energy.

d. The explanation of amplification using a tunnel diode as presented in paragraphs 3–9 through 3–10, therefore, is on a first-order basis. The use of the tunnel diode to achieve gain and oscillation is explained without con-

sidering the mathematical concept of negative resistance. The latter is discussed in paragraph 3–11.

3–8. Varying DC Resistance

a. Figure 3–6 shows the tunnel diode and its current-voltage chart with specific values of voltage applied and the resultant current flow. The current-voltage chart compared with the current-voltage chart in Fig. 2–19 is idealized into straight-line portions to simplify the explanation. This simplification does not change the validity of the presentation. Note that the portions of the chart (Fig. 3–6A) from zero (0) to peak tunnel current (1) and from valley current (3) to injection current (4) represent *constant*

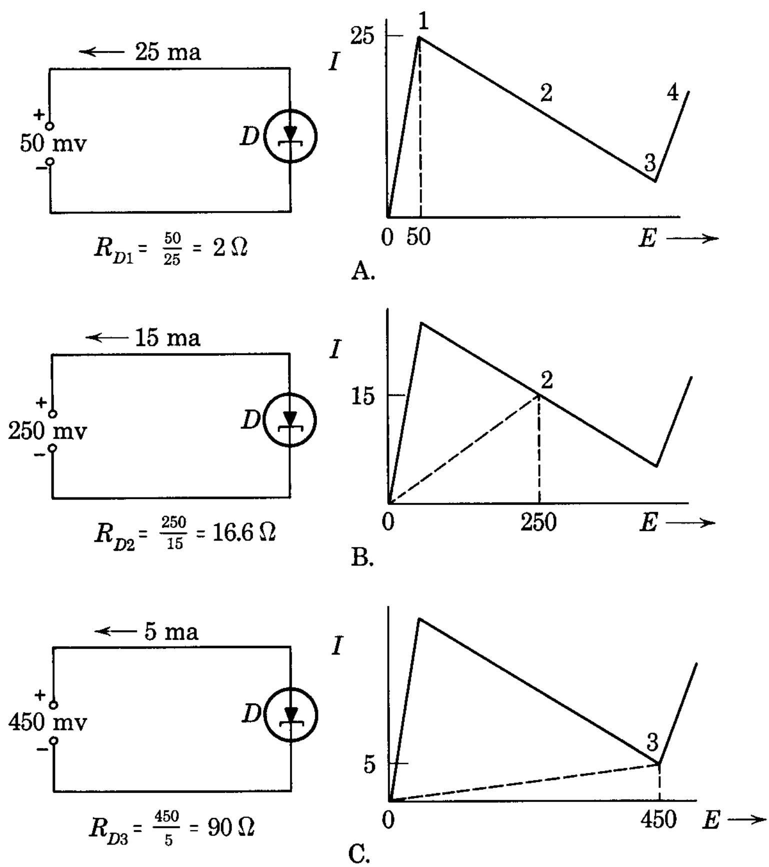


FIG. 3–6. Tunnel diode and current-voltage chart, showing three values of applied forward bias, and corresponding values of dc resistance

dc positive resistances because the current increases in a steady ratio with applied voltage. The portion of the chart from peak tunnel current (1) to valley current (3) represents an increasing dc resistance because the current decreases in a steady ratio with increasing applied voltage.

1. Figure 3–6A shows that, with 50 mv of applied forward bias, a current of 25 ma flows. Dc resistance at point (1) on the current-voltage chart is found by using Ohm's law. The resistance of the tunnel diode at this point is designated R_{D1} and equals 2 ohms.
2. Figure 3–6B shows, that with 250 mv of applied forward bias, a current of 15 ma flows. Dc resistance at point (2) again is found by using Ohm's law. The resistance of the tunnel diode at this point is designated R_{D2} and equals 16.6 ohms.
3. Figure 3–6C shows that, with 450 mv of applied forward bias, a current of 5 ma flows. Dc resistance at point (3) is found by dividing the total voltage by the total current. The resistance of the tunnel diode at this point is designated R_{D3} and equals 90 ohms.

b. In *a*1, *a*2, and *a*3 above, it has been shown that the dc resistance of the tunnel diode varies from a *minimum* value at its peak current (point 1, Fig. 3–6A) to a *maximum* value at its valley current (point 3). The resistance values have been calculated at only three points; these are sufficient for the subsequent discussions (par. 3–9 and 3–10). The specific values at any other points could just as easily be calculated. Note that the dc resistance calculated will be the same at each point as long as the magnitude of current flow through the device in the direction shown occurs. In other words, the dc resistance is an *instantaneous* value and applies no matter how brief or how long the period of current flow.

3–9. Tunnel Diode Series Amplifier

a. Figure 3–7 shows a series circuit which includes a battery (E_B), a voltage signal source (e_s), a tunnel diode, and a positive load resistor (R_L). For the period being considered, there is no output from the signal source. The current (I) measured is 15 ma in the direction indicated by the arrow. The total applied voltage (E_A) is the battery voltage of 1000 mv. By Ohm's law, it can be shown that the voltage drop (E_R) across resistor R_L is 750 mv, and the voltage (E_D) across the tunnel diode is 250 mv. Under these conditions, the tunnel diode has a positive resistance (R_{D2}) of 16.6 ohms (Fig. 3–6B). In Fig. 3–7, the levels of current and voltages of interest are shown by the individual graphs.

b. Figure 3–8 shows the same circuit at an instant of time when the

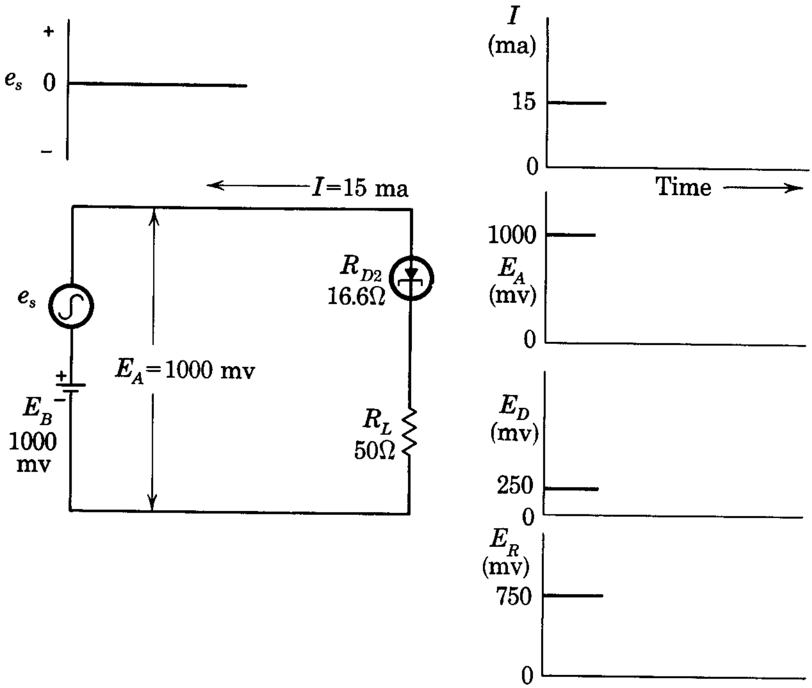


FIG. 3–7. Tunnel diode series amplifier, zero signal voltage applied

signal source (e_s) has reached a peak of 300 mv, series aiding the battery voltage (E_B). The applied voltage (E_A) is now 1300 mv. The current (I) is 25 ma. By Ohm's law, it can be shown that the voltage drop (E_R) across resistor R_L is now 1250 mv. This leaves a voltage (E_D) of 50 mv across the tunnel diode, the positive resistance of which is now 2 ohms (Fig. 3–6A).

c. Figure 3–9 represents the same circuit at an instant of time when the signal voltage (e_s) has reached a peak of 300 mv, opposing the battery voltage (E_B). The applied voltage is 700 mv. The current (I) is now 5 ma. The voltage drop (E_R) across resistor R_L must be 250 mv. This leaves a voltage (E_D) of 450 mv across the tunnel diode, the positive resistance of which is now 90 ohms (Fig. 3–6C). Note that the most important aspect of the action illustrated in Figs. 3–7, 3–8, and 3–9 is that the signal voltage (e_s) has had a peak-to-peak swing of 600 mv, and that the voltage (E_R) across load resistor R_L has had a peak-to-peak swing of 1000 mv, or an increase of 400 mv (1000 mv – 600 mv). Voltage gain has occurred (par. 3–10). In addition, the input ac signal voltage (e_s) is *in phase* with the output (ac) signal voltage; compare e_s with E_R in Fig. 3–9.

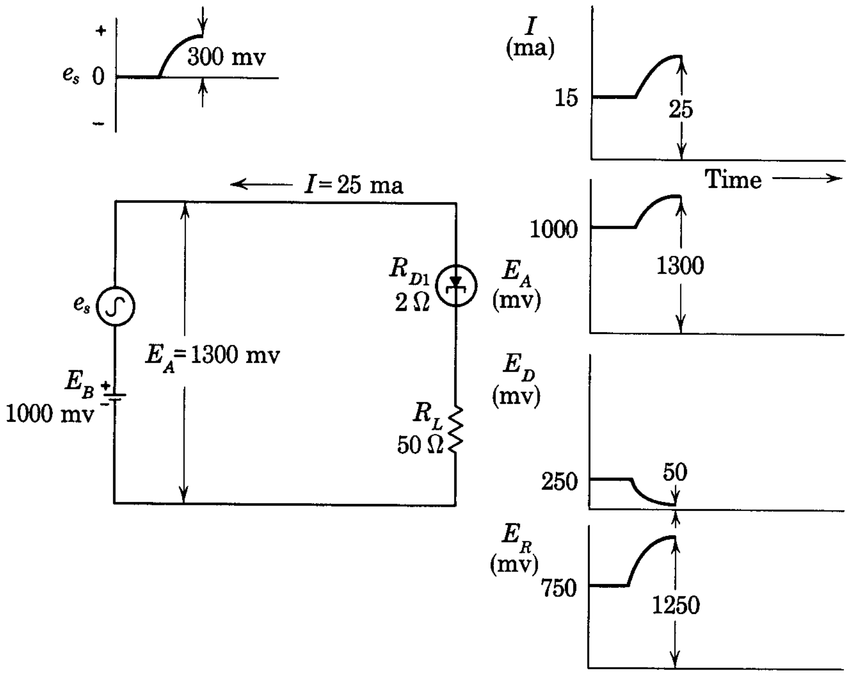


FIG. 3–8. Tunnel diode series amplifier, series-aiding signal voltage applied

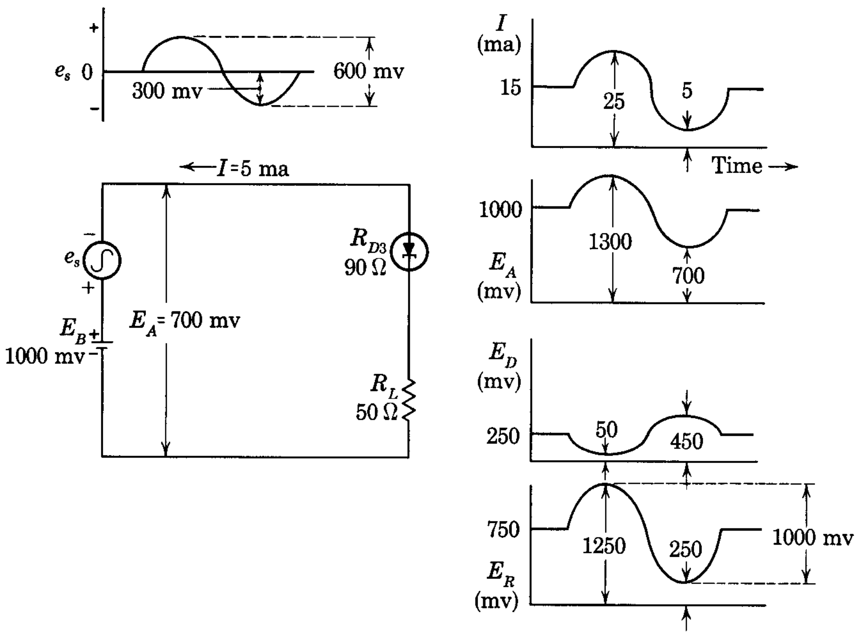


FIG. 3–9. Tunnel diode series amplifier, series-opposing signal voltage applied

3–10. Series Gain by Varying DC Resistance

a. The gain achieved (par. 3–9) by use of the tunnel diode can be explained on the basis of the varying instantaneous dc positive resistance of the tunnel diode. Consider Fig. 3–7 which actually shows a quiescent condition (no voltage signal applied). In this case the resistance of the tunnel diode is 16.6 ohms.

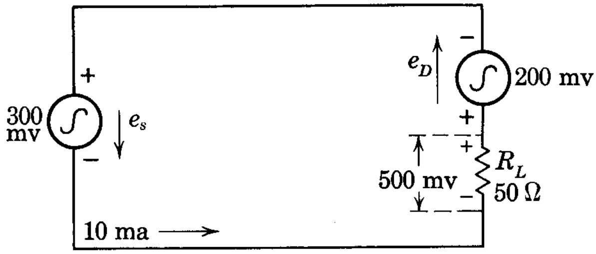
b. The series aiding signal voltage (Fig. 3–8) normally causes an increase in current. However, the dc resistance of the tunnel diode has *decreased* to 2 ohms so that the *total* increase in current is *greater* than would have occurred had the tunnel diode resistance remained at 16.6 ohms. The *greater* current increase has caused a voltage increase across R_L of 500 mv (1250 – 750) instead of *less* than 300 mv.

c. In Fig. 3–9 the series opposing signal voltage normally causes a decrease in signal current. However, the dc resistance of the tunnel diode has *increased* to 90 ohms, causing a *greater* decrease in current than that caused by the signal voltage alone. The *greater* decrease in current has caused a voltage decrease across resistor R_L of 500 mv (750 – 250) instead of *less* than 300 mv. *Therefore, it is the varying dc resistance of the tunnel diode, controlled by the signal source, that has shifted the dc voltage (and energy) of the battery in such a manner as to give ac voltage, current, and power gain in load resistor R_L.*

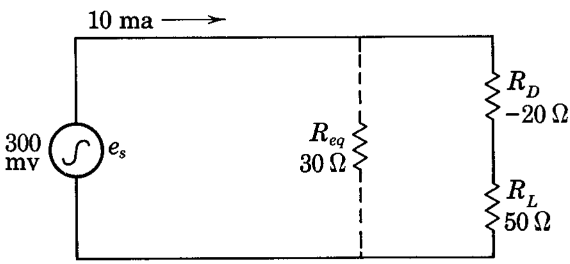
3–11. Series Amplifier, Equivalent Circuit

a. General. The equivalent circuit of an amplifier eliminates the dc biasing voltages and components of the basic circuit (Fig. 3–7) and includes only those components that directly affect the ac signal alone. All active circuit devices, such as electron tubes, crystals, transistors, and tunnel diodes, are replaced by circuit equivalents of generators, resistors, capacitances, etc. These circuit equivalents permit computations of voltage, current and power gain, cutoff frequency, input and output impedances, and other important quantities. The equivalent circuit is necessary because the formulas used for making computations are expressed only in terms of such quantities as current, voltage, resistance, inductance, and capacitance. The following discussion involves the equivalent circuit for the tunnel diode used as a series amplifier (par. 3–7 through 3–10). The amplifier is considered to be a very low frequency (1000 cps) amplifier and the equivalent circuit of the tunnel diode at this frequency is more simple than that at higher frequencies. For the equivalent circuit of the tunnel diode at high frequencies, refer to paragraph 3–17.

b. Signal Generator Equivalent. In Fig. 3–10A, which is a possible equivalent circuit of the amplifier shown in Fig. 3–7, the tunnel diode has



A. Tunnel diode replaced by in−phase generator



B. Tunnel diode replaced by negative resistance

FIG. 3–10. Equivalent circuits of tunnel diode amplifier

been replaced by an equivalent signal generator (e_D) of the same frequency as the signal source (e_s) and in phase. The voltage of the equivalent signal generator must be 200 mv to account for the total 500 mv drop across resistor R_L (par. 3–10). The total 500 mv ($e_s + e_D$) obviously is necessary to account for an ac current of 10 ma (par. 3–16). If only voltage e_s were in the circuit, then the ac current (i) would be:

$$i = \frac{300}{50}$$

$$= 6 \text{ ma}$$

This current would not reflect the fact that 10 ma of ac current flows through load resistor R_L. Note that in this equivalent circuit the tunnel diode *acts like an ac generator and a source of energy* (par. 3–7*b*). Representation of the tunnel diode as an ac generator is not extensively used, because the value of its output voltage depends upon the value of the signal voltage (e_s). The particular value of equivalent voltage (e_D), therefore, is not an

inherent characteristic (fixed value) of a particular tunnel diode. Furthermore, impedances *seen* by the signal source or the load are not readily calculated with this equivalent circuit.

c. Negative-Resistance Equivalent. Figure 3–10B shows an equivalent circuit of the amplifier of Fig. 3–7.

1. In this equivalent circuit, the tunnel diode has been replaced by a resistance (R_D) of negative (−) 20 ohms. The total resistance (R_{eq}) *seen* by the signal source is calculated as follows:

$$R_{eq} = \frac{\text{signal source voltage } (e_s)}{\text{total current } (i)}$$

Therefore,

$$R_{eq} = \frac{300 \text{ mv}}{10 \text{ ma}}$$

$$= 30 \text{ ohms}$$

This value of resistance is shown in dashed lines in Fig. 3–10B. The *effect* of the tunnel diode has been to *reduce* the load resistance (R_L) from 50 ohms (actual value) to 30 ohms (equivalent value). The tunnel diode then *acts like a resistance of −20 ohms (50 − 30 = 20).*

2. Notice that the equivalent ac negative resistance (−20 ohms) of this tunnel diode is *not* the difference between its minimum dc resistance of 2 ohms (par. 3–8) and its maximum dc resistance of 90 ohms. These dc resistance values are calculated from total instantaneous current and voltage values. The ac resistance is calculated only from the *changes* in current and voltage values over the operating range. The ac negative resistance of any tunnel diode can be calculated directly from its current-voltage chart. For this tunnel diode (Fig. 3–11) the total current swing is from 25 ma to 5 ma; the current change (ΔI) therefore is 20 ma (25 −5). The total voltage swing is from 50 mv to 450 mv; the voltage change (ΔE) therefore is 400 mv (450 − 50). The negative resistance ($-R_D$) of the tunnel diode is:

$$R_D = \frac{\Delta E}{\Delta I}$$

Therefore,

$$R_D = \frac{400 \text{ mv}}{20 \text{ ma}}$$

$$= 20 \text{ ohms}$$

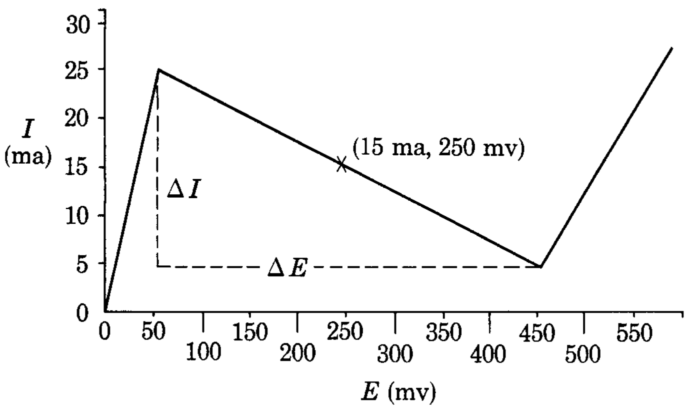


FIG. 3–11. Calculation of negative resistance of tunnel diode using its current-voltage chart

This is a *negative* resistance because current *decreases* with increasing voltage. *Note:* When a nonidealized current-voltage chart is used (Fig. 2–19), the negative resistance value calculated will be valid only if a straight-line portion of the curve is used.

3–12. Series Amplifier, Gain Formulas

a. Current Gain. The current gain (A_i) of a series amplifier may be defined as the ratio of the current (i_2) through the load resistor R_L with the tunnel diode in the circuit (Fig. 3–12B) to the current (i_1) through load resistor R_L without the tunnel diode in the circuit (Fig. 3–12A).

1. By definition the expression for current gain is:

$$A_i = \frac{i_2}{i_1}$$

2. The derived formula for current gain is:

$$A_i = \frac{R_L}{R_L - R_D}$$

Note: Refer to Appendix B for derivation of this formula. In this formula use the absolute (positive) numerical value for R_D. Note also that the current gain equals the *ratio of the impedance transformation* (R_L without the tunnel diode; $R_L - R_D$ with the tunnel diode).

3. The following are numerical examples of the use of this formula:

Example 1. Given: $R_D = 20$ ohms
$R_L = 50$ ohms

Find A_i by substituting numerical values (2 above):

$$A_i = \frac{50}{50 - 20}$$

$$A_i = \frac{50}{30}$$

$$A_i = 1.66$$

Example 2. Given: $R_D = 20$ ohms
$R_L = 25$ ohms

Find A_i by substituting values:

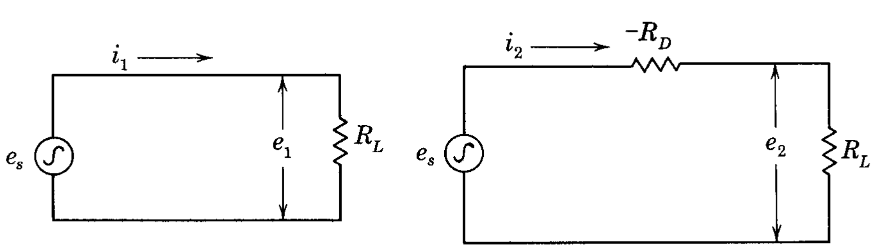
$$A_i = \frac{25}{25 - 20}$$

$$A_i = \frac{25}{5}$$

$$A_i = 5$$

4. Note that the gain is larger (3 above) when load resistor R_L value approaches the value of R_D. Inspection of the gain formula (2 above) indicates that a theoretical gain of infinity is possible by making $R_L = R_D$; i.e., the denominator equals zero. Actually an unstable (oscillatory) condition occurs (par. 3–15).

b. Voltage Gain. The voltage gain (A_v) is defined as the ratio of output voltage (e_2) with the tunnel diode in the circuit (Fig. 3–12B) to the output voltage (e_1) without the tunnel diode in the circuit (Fig. 3–12A).



A. Signal source and load resistor

B. Equivalent circuit, series tunnel diode amplifier

FIG. 3–12. Current and voltage designation for source and load-resistor circuit and for equivalent series circuit using tunnel diode

1. The expression for voltage gain is:

$$A_v = \frac{e_2}{e_1}$$

2. The derivation of A_v formula is in Appendix B. It is found to be identical to the formula for current gain:

$$A_v = \frac{R_L}{R_L - R_D}$$

3. The numerical values calculated for current gain in the examples (*a*3 above) apply also to voltage gain.

c. Power Gain. The power gain (G) is the product of current gain (A_i) and power gain (A_v).

1. The power gain formula is:

$$G = \left(\frac{R_L}{R_L - R_D}\right)^2$$

or

$$= A_i A_v$$

or

$$= A_i^2$$

or

$$= A_v^2$$

2. Examples using the corresponding information in *a*3 above are:

Example 1. Given current gain (A_i) of 1.66, find power gain:

$$G = A_i^2$$
$$G = (1.66)^2$$
$$G = 2.76$$

Example 2. Given current gain (A_i) of 5, find power gain:

$$G = A_i^2$$
$$G = 5^2$$
$$G = 25$$

3–13. Series Amplifier, Graphical Analysis

Graphical analysis can be used to determine voltage gain in a series tunnel diode amplifier. The analysis is performed by adding the voltage drops across the diode and the load resistor to yield a voltage curve of the total applied voltage. The input signal voltage is projected onto the total applied voltage curve and its effect noted on the load-resistor curve.

a. Figure 3–13A shows a voltage-current chart for a circuit such as that shown in Fig. 3–7. However, values are different for the applied battery voltage, signal voltage, and load resistor. Note that in Fig. 3–13A the voltage is plotted vertically to simplify voltage additions; the same technique can be used by adding voltages horizontally. Three curves are shown: curve *OABC* which is the idealized tunnel diode curve used previously; curve *OD* which shows the voltage drop across a load resistor of 25 ohms; and curve *OEFG* which is the sum of the other two curves.

1. The true tunnel diode curve is normally supplied by the manufacturer.

2. The load-resistor curve (a straight line) is determined by two points: one point is the zero current-voltage point; the second point is determined by arbitrarily choosing a current (e.g., 25 ma) and determining the voltage across the resistor (25 ohms) with this current. The second point, therefore, is 25 ma and 625 mv (25 ma $\times$ 25 ohms = 625 mv). A straight line from zero to this point determines the load-resistor voltage line.

3. The total applied voltage is the sum of the two curves 1 and 2 above. This is obtained by: adding *OA* to *OD* to give *OE*; adding *AB* to *OD* to give *FE*; and adding *BC* to the corresponding portion of *OA* to give *FG*.

b. To avoid congestion, Fig. 3–13B repeats only those portions of the curves that are of interest in the graphical analysis. These portions are the load-resistor curve (*OD*), the negative-resistance portion (*AB*) of the tunnel diode curve, and their sum (*FE*).

1. The quiescent point (15 ma, 625 mv) for the amplifier is marked by a dot on the applied voltage curve.

2. Assume an input signal of 50 mv peak-to-peak. Project this signal onto the applied voltage curve (each side of the quiescent point) as shown. The projections intersect the latter curve at 600 and 650 mv.

3. The intersections on the applied voltage curve are projected vertically down until the load-resistor curve (*OD*) is intersected. The latter intersections are projected horizontally to the left to produce the output voltage signal. Note that the intersections occur at 250 mv and 500 mv. The output signal, unchanged in phase, is 250 mv (500 mv − 250 mv) peak-to-peak.

c. The voltage gain (A_v) is the ratio of output voltage (250 mv) to input voltage (50 mv). A_v = 250 mv $\div$ 50 mv; the voltage gain is 5. Note that this value agrees with that calculated by using the gain formula (par. 3–12). In that case the same tunnel diode was used with the same load resistor.

3–14. Series Amplifier, Gain-Formula References

Analysis of Fig. 3–13 shows that the same current flows through the load resistor that is delivered by the signal source. Projections of the voltage

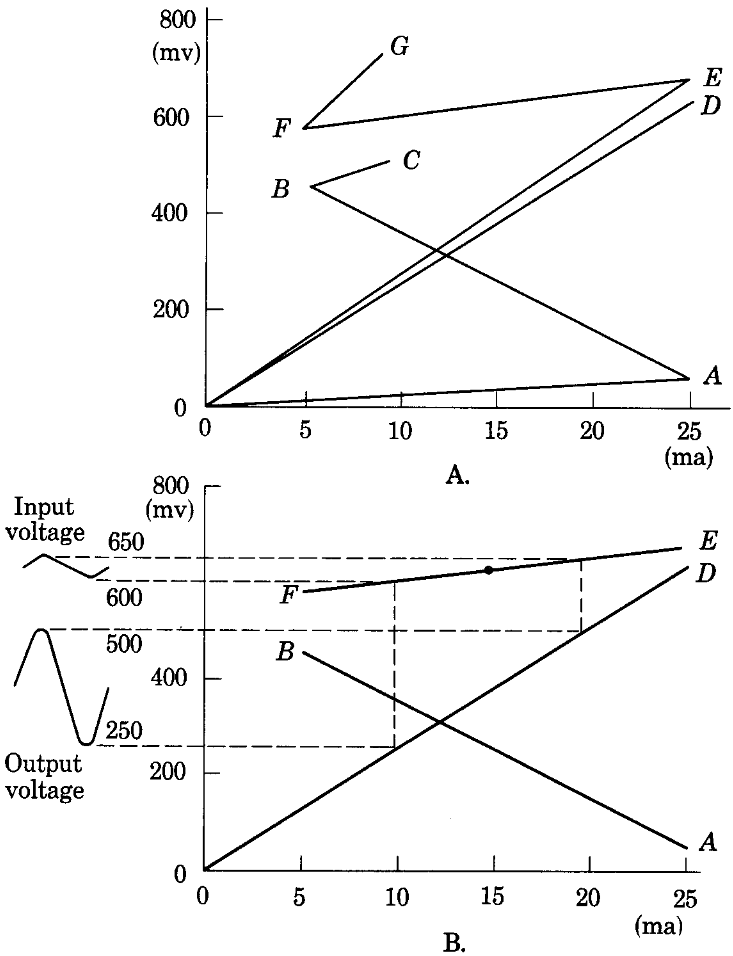


FIG. 3–13. Voltage-current charts of tunnel diode, load resistor, and voltage sum for graphical analysis of series amplifier

signal from the total voltage curve and from the load-resistor curve vertically down to the current axis gives the same current magnitude of 15 ma (20 ma − 5 ma). In other words, there appears to be no current gain, or the current gain is one. This observation appears to contradict the data in paragraph 3–12*a* in which current gain in a simple series amplifier is indicated. In many publications writers have also made the statement that the current gain is *one* in a series amplifier. Both the data in paragraph 3–12*a* on current gain and the latter statement are true. If current gain is arbitrarily defined as the ratio of output current with the tunnel diode in

the circuit to output current without the tunnel diode in the circuit (par. 3–12a), then there is current gain; if current gain is arbitrarily defined as the ratio of input current (from the signal source) to output current (in the load) only with the tunnel diode in the circuit, then current gain is *one*. The tendency in the field is toward the latter definition. Using the latter definition, the formula for power gain (par. 3–12) is also affected; in either case power gain (G) is the product of current gain (A_i) and voltage gain (A_v). The formulas for current gain and power gain used in the literature will differ, depending on the particular definition referenced. Table 3–1 lists

TABLE 3–1

Gain	Two Circuits	One Circuit
Voltage (A_v)	$\dfrac{R_L}{R_L - R_D}$	$\dfrac{R_L}{R_L - R_D}$
Current (A_i)	$\dfrac{R_L}{R_L - R_D}$	1
Power (G)	$\left(\dfrac{R_L}{R_L - R_D}\right)^2$	$\dfrac{R_L}{R_L - R_D}$

the applicable formulas. The first column lists the type of gain being considered; the second column (*Two Circuits*) lists the formulas used when gain is considered a comparison of values in circuits with and without the tunnel diode; the third column (*One Circuit*) lists the formulas when gain is considered a ratio of output (in the load) to input (from the signal source) only when the tunnel diode is in the circuit.

SECTION III. OSCILLATION, SERIES ARRANGEMENT

3–15. Series Oscillator

a. The formula for current gain in a series amplifier (par. 3–12) was determined to be as follows:

$$A_i = \frac{R_L}{R_L - R_D}$$

b. In this formula, R_L is the load resistance, and R_D is the negative resistance of the tunnel diode. If these two values were equal, the current gain (A_i) would be infinite. Actually, it is found that the circuit breaks into oscillation (instability) and cannot function as an ordinary amplifier.

c. Figure 3–14A shows the idealized current-voltage chart of the tunnel diode, the negative resistance of which has been determined to be −20 ohms (par. 3–11). Figure 3–14B shows this tunnel diode in a series circuit with a load resistor (R_L) equal to 20 ohms, and an applied voltage of 550 mv. In Fig. 3–14A the 20-ohm load line is drawn on the current-voltage chart. The load line is determined by two points.

1. If the current through the tunnel diode is zero, then the voltage across the load resistor equals the applied voltage of 550 mv. Therefore, the point of zero current and 550 mv is one point on the 20-ohm load line.
2. If the voltage across the tunnel diode is zero, then the applied voltage is entirely across the load resistor; the current flow is: 550 mv ÷ 20 ohms = 27.5 ma. Therefore the second point on the load line is 27.5 ma for zero voltage (across the tunnel diode).
3. A straight line is drawn between the two points. Note that the load line in heavy dash lines coincides with the negative resistance portion of the tunnel diode curve.

d. In Figs. 3–14B, C, and D, the tunnel diode dc resistance at three points on its current-voltage chart is shown. These resistance values (R_{D1}, R_{D2}, and R_{D3}) were previously determined (par. 3–8). Actually any other num-

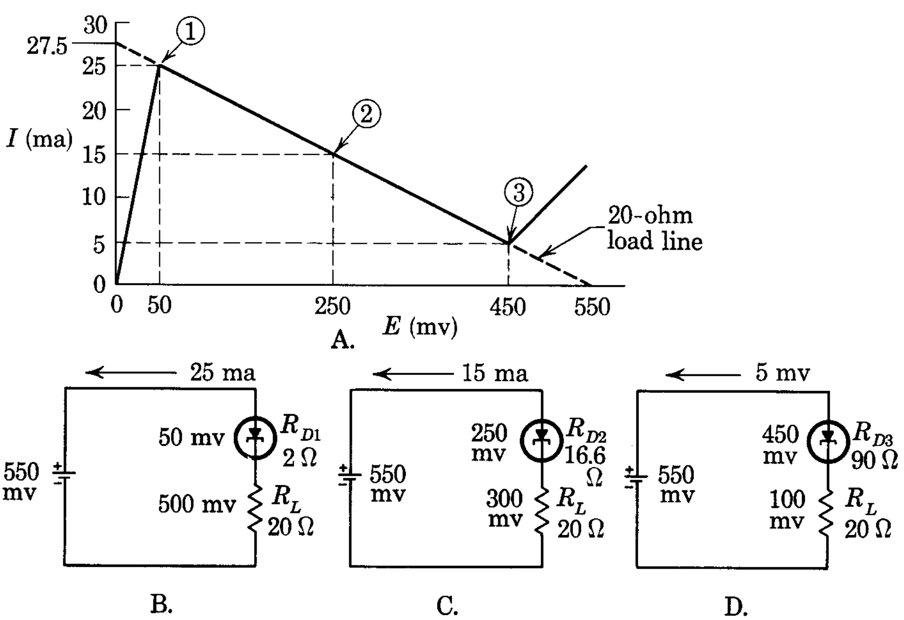


FIG. 3–14. Current-voltage chart with superimposed load line and circuits showing instantaneous voltage distributions

ber of points on the negative-resistance portion might have been used and similar circuits drawn. At which point will the circuit stabilize? None! The current flow will vary from point 1 to point 3 at a given rate. The given rate could be determined by distributed inductance and capacitance (par. 3–17), or a tuned circuit could be placed in series with the tunnel diode; the resonant frequency of the tuned circuit would be the major factor in determining the rate at which the operating point would move up and down the *negative resistance* portion of the tunnel diode current-voltage chart. Obviously, the circuit is an oscillator.

3–16. Oscillation and Power Shift Relations

a. In the study of amplification by use of the *negative resistance* of the tunnel diode it was shown that amplification (par. 3–10) of the applied signal was brought about by the varying dc resistance of the tunnel diode. The reason for the *start* of the oscillation in a series amplifier when the load resistance equals the negative resistance can be understood by studying the power shift relations.

b. Consider Fig. 3–11. Assuming the quiescent point marked × (15 ma, 250 mv), the maximum current (I) is 10 ma (one half of a 20 ma peak-to-peak swing). The negative resistance (R_D) of the diode is 20 ohms (par. 3–11). The maximum power (P) that can be *shifted* by the tunnel diode can be determined as follows:

1. Write formula for power:

$$P = I^2R$$

2. Substitute maximum current and negative resistance values:

$$P = (0.01)^2 20$$

Note: 10 ma equal 0.01 amp. Therefore

$$P = 2 \text{ mw}$$

c. The peak power (*b*2 above) that can be shifted will be compared to the power absorbed by various possible load resistors. Table 3–2 lists various load resistors, and the power absorbed if a current of 10 ma passes through the resistor.

d. Table 3–2 shows that series load resistors of 50, 40, or 30 ohms absorb more power than the tunnel diode can *shift* (2 mw) at an ac rate from the battery. As a result, oscillations cannot be sustained and will be quickly damped for lack of power. At 20 ohms (the same value as that of the negative resistance) and lesser values, the tunnel diode can shift sufficient energy to sustain oscillations.

TABLE 3–2

Load Resistor (ohms)	Power Absorbed (mw)
50	5
40	4
30	3
20	2
15	1.5
10	1
5	.5

e. It can be stated, therefore, that any tunnel diode used in a series circuit will sustain oscillations if the series load resistance is *equal to or is less than the negative resistance of the tunnel diode.* Because oscillations cannot be sustained at resistance values *higher* than the negative resistance when in a series arrangement, the circuit is referred to as *open-circuit stable.*

3–17. Tunnel Diode Equivalent Circuit

a. General. Figure 3–15 shows the complete equivalent circuit of the tunnel diode for use at high frequencies. Inductor L_s and resistor R_s represent parasitic elements introduced by connecting leads. Capacitor C_D represents capacitance caused mainly by the internal depletion region; resistor R_D represents the ac negative resistance of the device. The tunnel diode equivalent circuit is covered in this section because it acts like a series-resonant circuit when its input terminals are ac short-circuited. The maximum frequency of oscillation of a tunnel diode may be determined by its self-resonant frequency (*b* below), or its resistive cutoff frequency (*c* below).

b. Self-resonant Frequency. The self-resonant frequency (f_s) of the

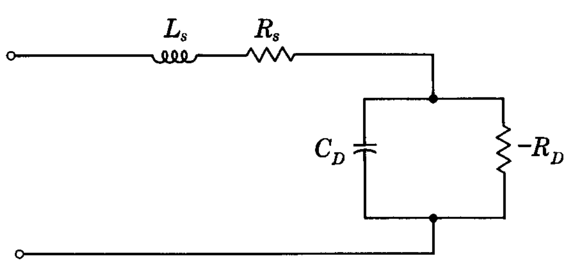


FIG. 3–15. High-frequency equivalent circuit of tunnel diode

tunnel diode is the frequency of resonance of the inductive reactance and the capacitive reactance.

1. The formula for calculating f_s is derived in Appendix B and is:

$$f_s = \frac{1}{2\pi C_D R_D}\sqrt{\frac{C_D R_D{}^2 - L_s}{L_s}}$$

2. Typical values of these parameters for tunnel diodes having 1/8 inch leads are:

(a) C_D, 5 to 10 mmf.
(b) R_D, 50 to 120 ohms.
(c) L_s, 5 to 10 mμh (millimicrohenries; 10^{-9} henry).
(d) R_s, approximately 1 ohm (not used to calculate f_s).

3. It can be shown by using the formula (in 1 above) that a tunnel diode having $C_D = 7$ mmf, $R_D = 100$ ohms, $L_s = 6$ mμh, and $R_s = 1$ ohm, will have a self-resonant frequency (f_s) of 750 kmc. Refer to Appendix B for sample calculations.

c. *Resistance Cutoff Frequency.* The resistance cutoff frequency (f_r) of a tunnel diode is the frequency at which the net negative resistance of the (equivalent) circuit goes to zero and oscillation ceases.

1. The formula for f_r is derived in Appendix B and is:

$$f_r = \frac{1}{2\pi C_D R_D}\sqrt{\frac{R_D - R_s}{R_s}}$$

2. Normally it would be expected that the net negative resistance of the tunnel diode equivalent circuit is independent of frequency and is simply the difference of R_s and R_D. Actually the above formula indicates the *frequency* at which the power absorbed by the positive resistance (R_s) is equal to the power that the negative resistance is capable of *shifting* from the dc battery to make up for the ac losses. This power is frequency-dependent because of the shunting effect of capacitor C_D on resistor R_D (par. 3–18). In other words, the ac current that flows through resistor R_s at the cutoff frequency is much larger than the ac current flow through resistor R_D.

3. Note that if the resistance cutoff frequency (f_r) for a given unit is lower than the self-resonant frequency (f_s), the unit will not resonate at its calculated self-resonant frequency. Units having parameters indicated in *b*2 above have resistance cutoff frequencies as high as 10 kmc. Some units have been made to oscillate at frequencies as high as 100 kmc, indicat-

ing an f_r greater than 100 kmc. For most units f_r is in the 1- to 3-kmc range. However, if R_s is increased to 50 ohms by load and generator resistance, the resistance cutoff frequency will be reduced by a factor of 10.

3–18. Frequency Dependence of Negative Resistance

a. The effective negative resistance (R_D') displayed by a given tunnel diode in a circuit depends on the operating frequency (f). The negative resistance (R_D) calculated from the current-voltage chart by using incremental dc values (par. 3–11) does not consider the shunting effect of the inherent capacitance (C_D) of the diode. The following formula (derived in par. B-3 of Appendix B) permits calculation of the effective negative resistance at any particular operating frequency:

$$R_D' = \frac{R_D}{(2\pi f C_D R_D)^2 + 1}$$

Numerical examples of the use of this formula are given in *b* below.

b. Assume a tunnel diode having a negative resistance (R_D) of 100 ohms and an inherent capacitance (C_D) of 10 mmf.

Example 1. Calculate R_D' at an operating frequency (f) of 100 mc (10^8 cps):

1. Substitute known values in the formula (*a* above)

$$R_D' = \frac{100}{(2 \times 3.14 \times 10^8 \times 10 \times 10^{-12} \times 100)^2 + 1}$$

2. Determine value:

$$R_D' = 71.4 \text{ ohms}$$

Example 2. Calculate R_D' at an operating frequency (f) of 1 kmc (10^9 cps):

1. Substitute known values in the formula (*a* above):

$$R_D' = \frac{100}{(2 \times 3.14 \times 10^9 \times 10 \times 10^{-12} \times 100)^2 + 1}$$

2. Determine value:

$$R_D' = 2.47 \text{ ohms}$$

The above examples indicate that the effective negative resistance of this tunnel diode having $R_D = 100$ ohms (at audio frequencies) is reduced to 71.4 ohms at 100 mc, and only 2.47 ohms at 1 kmc.

SECTION IV. AMPLIFICATION, PARALLEL ARRANGEMENT

3–19. General

Amplification can be achieved by placing a tunnel diode in *parallel* with a load resistor. As in the case of the series amplifier (par. 3–9), amplifica-

tion by a parallel amplifier will first be investigated by considering the instantaneous unidirectional currents in the circuit (par. 3–20 and 3–21). Subsequently, the equivalent circuit of the tunnel diode embodying the second-order concept of negative resistance (par. 3–11) is developed (par. 3–22). Gain formulas derived from the equivalent circuit are presented and interpreted in paragraph 3–23.

3–20. Tunnel Diode Parallel Amplifier

Figure 3–16 shows a tunnel diode parallel amplifier consisting of a battery (E_B), a signal source (e_s), and a load resistor (R_L) in parallel with a tunnel diode. The tunnel diode being used here has the same current-voltage chart analyzed in paragraph 3–8 and shown in Fig. 3–6. It has been determined that its resistance at peak current is 2 ohms (R_{D1}); midway between peak and valley currents the resistance is 16.6 ohms (R_{D2}); and at the valley current it is 90 ohms (R_{D3}).

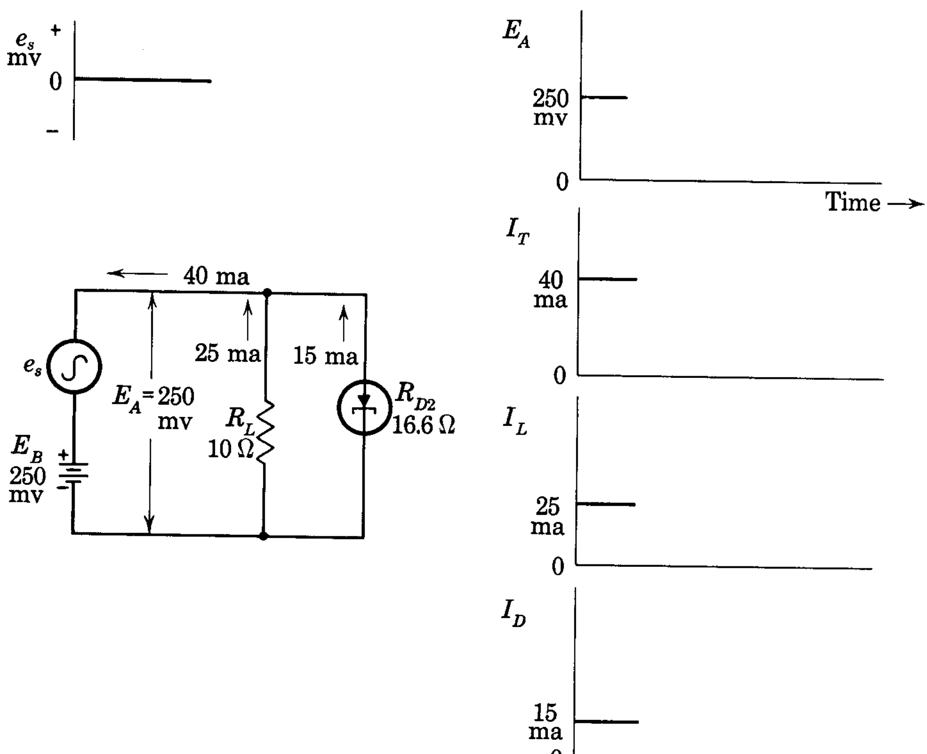


FIG. 3–16. Tunnel diode parallel amplifier, zero signal voltage applied

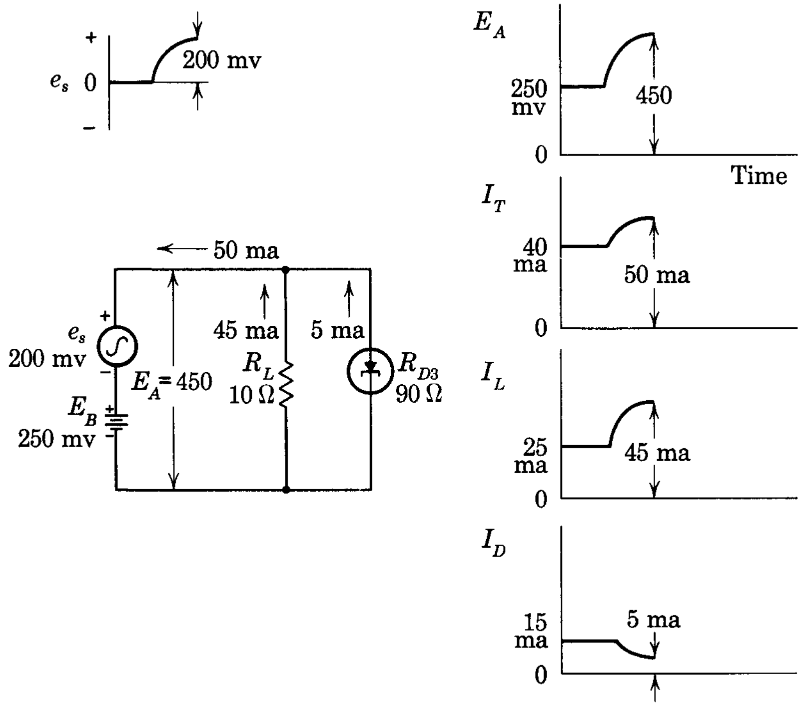


FIG. 3–17. Tunnel diode parallel amplifier, signal voltage and biasing battery series aiding

a. In Fig. 3–16, the quiescent condition is shown; i.e., the signal voltage is zero, and the applied voltage (E_A) equals the battery voltage (250 mv). It can be shown by Ohm's law that the current (I_L) through resistor R_L is 25 ma in the direction shown, and the current (I_D) through the tunnel diode is 15 ma in the direction shown. The total current (I_T) is the sum of I_L and I_T or 40 ma (25 ma + 15 ma). The applied voltage and the three currents are indicated in the separate charts.

b. Figure 3–17 shows the circuit at the instant when the signal voltage (e_s) is series aiding the battery voltage (E_B). The total applied voltage (E_A) is now 450 mv. With this applied voltage, the tunnel diode draws 5 ma, and resistor R_L draws 45 ma. The total current (I_T) is now 50 ma. All currents flow in the direction shown. The applied voltage and the three currents at this instant are indicated in the separate charts.

c. Figure 3–18 shows the circuit at the instant when the signal voltage (e_s) is series opposing the battery voltage (E_B). The total applied voltage

(E_A) is now 50 mv. With this applied voltage the tunnel diode draws 25 ma, and resistor R_L draws 5 ma. The total current (I_T) is now 30 ma. All currents flow in the direction shown. The applied voltage and the three currents are shown in separate charts.

d. Note that the total current (see I_T chart in Fig. 3–18) has varied from 50 ma to 30 ma, or 20 ma peak-to-peak. Note also that the peak-to-peak

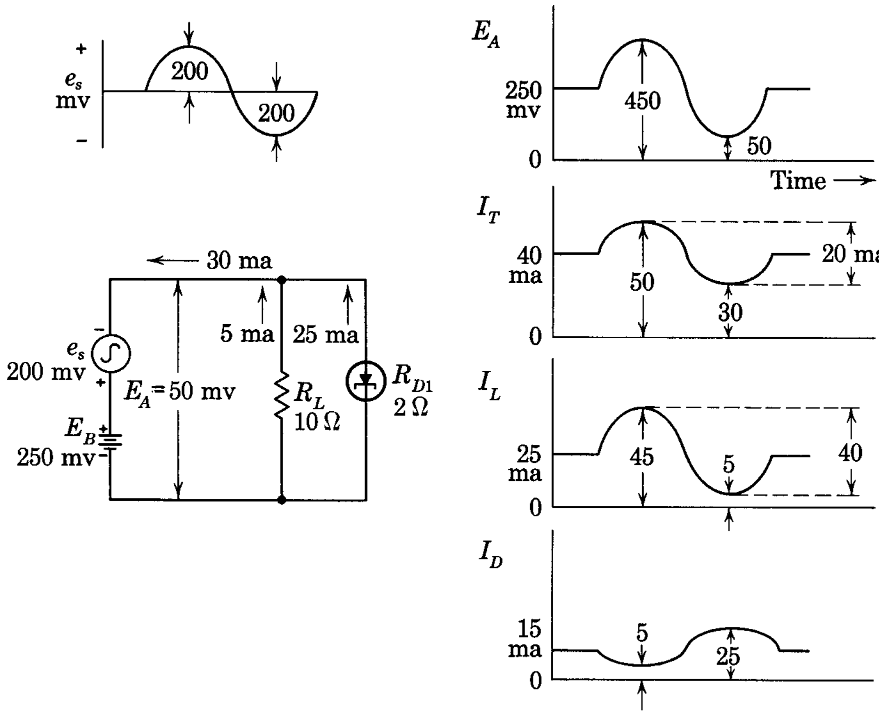


FIG. 3–18. Tunnel diode parallel amplifier, signal voltage and biasing battery series opposing

current variation through resistor R_L is 40 ma (see I_L chart). With the signal source causing a total current variation of 20 ma and the output current (I_L) equal to 40 ma, there has been a *current gain* of two. This current gain is caused by the varying dc resistance of the tunnel diode (par. 3–21).

e. Because of the parallel arrangement, there has been no voltage gain. Note that the peak-to-peak voltage swing across resistor R_L is 400 mv (450 mv − 50 mv). Note also that the peak-to-peak voltage swing of the signal voltage (E_S) is also 400 mv (200 mv positive and 200 mv negative). In addition, there has been no phase reversal between the total input (ac)

current and the output (ac) current; compare the waveforms for I_T and I_L in their respective charts.

3–21. Parallel Gain by Varying DC Resistance

The current gain of two in the parallel amplifier (Fig. 3–18) and noted in paragraph 3–20*d* is caused entirely by the varying dc resistance of the tunnel diode.

a. Increased Applied Voltage. 1. When the applied voltage increases because of signal voltage e_s, the dc resistance of the tunnel diode rises from 16.6 ohms (at quiescence) to 90 ohms. As a result, instead of drawing *more* current, the tunnel diode draws 10 ma *less* current.

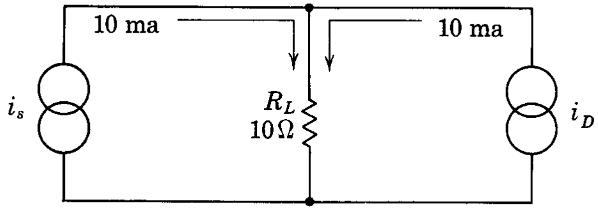
2. With the increased applied voltage, the load resistor draws 20 ma more current than that drawn at quiescence. The total added current that must be supplied by the signal source is only 10 ma [the difference between tunnel current (*a* above) and load resistor current]. Since an ac increase of 20 ma has occurred in the load resistor with only 10 ma supplied by the signal source, then 10 ma of battery current has been *shifted* from the tunnel diode to the load resistor by the varying dc resistance of the tunnel diode as *controlled* by the signal input voltage.

b. Decreased Applied Voltage. When the total applied voltage decreases because of signal voltage e_s, the dc resistance of the tunnel diode falls from 16.6 ohms (at quiescence) to 2 ohms. As a result, instead of drawing *less* current, the tunnel diode draws 10 ma more current. With decreased applied voltage the load resistor draws 20 ma *less* current than that drawn at quiescence. Without the tunnel diode in the circuit, the signal source would have to supply 20 ma of opposing current to the load resistor to reduce its current from 25 ma at quiescence to 5 ma at this instant. However, the tunnel diode *diverts* (or absorbs) 10 ma of the load current, so that the signal source need supply only 10 ma of opposing current. The overall action then is that the signal source supplied a 20 ma peak-to-peak current, while the current in the load has a peak-to-peak variation of 40 ma. The extra 20 ma in the load was *shifted* from the battery source by the dc resistance variation of the tunnel diode.

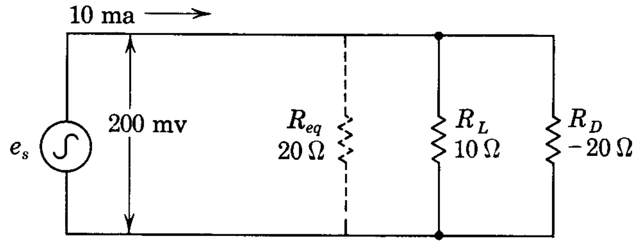
3–22. Parallel Amplifier, Equivalent Circuit

a. Current Generator Equivalent. A possible equivalent circuit for the parallel amplifier discussed in paragraph 3–21 is shown in Fig. 3–19A. The tunnel diode has been replaced by an equivalent current generator (i_D) of the same frequency as the signal source (i_s) and in phase. Note that for simplification the original voltage signal source (e_s) (Fig. 3–18) has been replaced by its equivalent current generator (i_s). The output current (i_D) of the tunnel diode equivalent circuit must be 10 ma to account for the

total current of 20 ma through load resistor R_L; only 10 ma is supplied by i_s. Note that in this equivalent circuit the tunnel diode *acts like an ac current generator and a source of energy* (par. 3–7b). Representation of the tunnel diode as an ac current generator is not extensively used, because the value of its output current depends upon the value of the signal current (i_s). The particular value of equivalent current (i_D) therefore is not an in-



A. Tunnel diode replaced by in-phase generator



B. Tunnel diode replaced by negative resistance

FIG. 3–19. Equivalent circuits of tunnel diode parallel amplifier

herent characteristic (fixed value) of a particular tunnel diode. Furthermore, impedances *seen* by the signal source or the load are not readily calculated with this equivalent circuit.

b. Negative-Resistance Equivalent. Figure 3–19B shows an equivalent circuit of the amplifier of Fig. 3–18.

1. In this equivalent circuit, the tunnel diode has been replaced by a resistance (R_D) of negative (−) 20 ohms. The total resistance (R_{eq}) *seen* by the signal source (e_s) is calculated as follows:

$$R_{eq} = \frac{\text{signal source voltage } (e_s)}{\text{total current } (i_s)}$$

Therefore,

$$R_{eq} = \frac{200 \text{ mv}}{10 \text{ ma}}$$

$$= 20 \text{ ohms}$$

This value of resistance is shown as dashed lines in Fig. 3–19B.

2. Note that the equivalent resistance (20 ohms) *seen* by the signal source is 10 ohms *greater* than load resistor R_L which is 10 ohms. No resistor having a *positive* resistance value can be placed across load resistor R_L so that the parallel combination has a greater resistance than R_L. Only a *negative* resistance can achieve this. The most simple method of determining the negative-resistance value of R_D in this circuit is to convert all resistance values to *conductance* values. Thus:

(a) Find total equivalent conductance:

$$G_{eq} = \frac{1}{R_{eq}}$$

$$= \frac{1}{20 \text{ ohms}}$$

$$= 0.05 \text{ mho}$$

(b) Find load conductance:

$$G_L = \frac{1}{R_L}$$

$$= \frac{1}{10 \text{ ohms}}$$

$$= 0.1 \text{ mho}$$

(c) Find tunnel diode conductance:

$$G_D = \frac{1}{R_D}$$

(d) Because conductances in parallel add:

$$G_{eq} = G_L + G_D$$

(e) Substitute known values (a) and (b) above:

$$0.05 = 0.1 + G_D$$

(f) Solve for G_D:

$$G_D = -0.05$$

(g) Convert G_D to R_D:

$$R_D = \frac{1}{G_D}$$

$$= \frac{1}{-0.05 \text{ mho}}$$

$$R_D = -20 \text{ ohms}$$

Note: In this case the tunnel diode *acts like a negative resistance of 20 ohms.* The negative resistance of a particular tunnel diode is calculated by using its current-voltage chart (par. 3–11c2). Note also that with the negative resistance known, the equivalent resistance calculated in 1 above can be verified as follows:

$$R_{eq} = \frac{(10)(-20)}{10 + (-20)} = \frac{-200}{-10} = 20$$

3–23. Parallel Amplifier, Gain Formulas

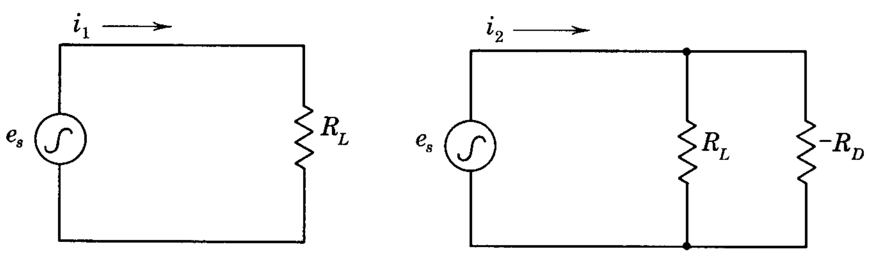
a. Current Gain. The current gain (A_i) of a parallel amplifier is the ratio of the current (i_1), drawn from the signal source with no tunnel diode present (Fig. 3–20A), to the current (i_2) drawn with the tunnel diode present (Fig. 3–20B).

1. By definition the expression for current gain is:

$$A_i = \frac{i_1}{i_2}$$

2. The derived formula for current gain is:

$$A_i = \frac{R_D}{R_D - R_L}$$



A. Signal source and load resistor

B. Equivalent circuit, parallel tunnel diode amplifier

FIG. 3–20. Current designations for signal source and load-resistor circuit, and for equivalent parallel circuit using tunnel diode

Note: Refer to Appendix B for the derivation of this formula. In this formula use the absolute (positive) numerical value for R_D.

3. The following are numerical examples of the use of this formula:

Example 1. Given: $R_D = 20$ ohms
$R_L = 10$ ohms

Find A_i by substituting numerical values (2 above):

$$A_i = \frac{20}{20 - 10} = \frac{20}{10} = 2$$

Example 2. Given: $R_D = 20$ ohms
$R_L = 15$ ohms

Find A_i by substituting values (2 above):

$$A_i = \frac{20}{20 - 15} = 4$$

4. Note that the gain is larger (3 above) when load-resistor R_L value approaches the value of R_D. If R_D equals R_L, a theoretical gain of infinity is possible because the denominator equals zero. Actually an unstable (oscillatory) condition occurs (par. 3–26).

b. Voltage Gain. The voltage gain (A_v) in a parallel arrangement always equals one (par. 3–20*e*); i.e., the same ac voltage appears across the load resistor, with or without the tunnel diode present. This statement, of course, assumes that there is no internal signal source resistance. If there is signal source resistance, then voltage gain as well as current gain can occur. This condition is covered in paragraph 3–25.

c. Power Gain. Power gain (G) is the product of voltage gain (A_v) and current gain (A_i). Since voltage gain in a parallel arrangement equals one, then power gain equals current gain. The formula and the numerical examples (*a* above) apply equally for power gain.

3–24. Parallel Amplifier, Graphical Analysis

Graphical analysis can be used to determine current gain in a parallel tunnel diode amplifier. The analysis is performed by adding currents through the diode and the load resistor to yield a current curve of the total current drawn from the signal source and battery. The input ac signal current is projected onto the total current curve and its effect noted on the load resistor curve.

a. Figure 3–21 shows the three required curves.

1. Curve *OA* is the idealized tunnel diode curve.

2. In this example, the load resistor is 15 ohms. One required point is the zero current, zero voltage point. Assuming 600 mv across the load, the current would be 40 ma (600 mv ÷ 15 ohms). Using the zero point and the 600 mv – 40 ma point, a straight line is drawn which represents the current through the load for specific voltage values.

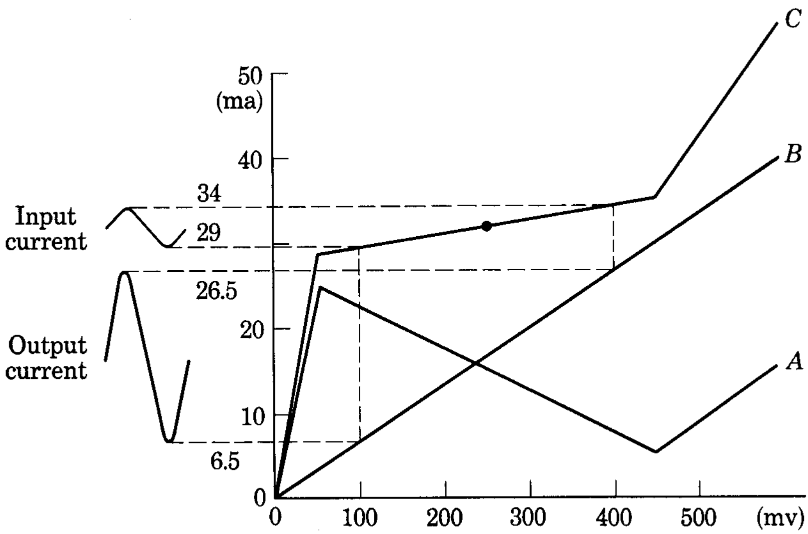


FIG. 3–21. Current-voltage charts of tunnel diode, load resistor and current sum for graphical analysis of parallel amplifier

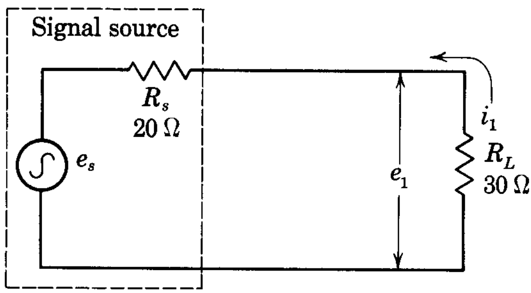
3. Curve *OC* is the sum of curves *OA* and *OB* on a point by point basis and represents the total current plotted against total applied voltage.

b. The quiescent point (31.5 ma, 250 mv) for the amplifier is marked on curve *OC* by a dot.

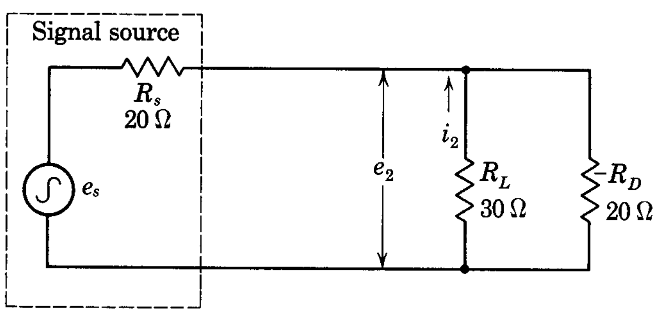
1. Assume an input signal current of 5 ma. peak-to-peak. Project this signal onto the total current curve (each side of the quiescent point) as shown. The projections intersect the total current curve at 29 and 34 ma.

2. The intersections on the total current curve are projected vertically down until the load-resistor curve is intersected. The latter intersections are projected horizontally to the left to produce the output current signal. Note that the intersections occur at 6.5 and 26.5 ma. The output current, unchanged in phase, is 20 ma peak-to-peak.

c. The current gain (A_i) is the ratio of output current (20 ma) to input current (5 ma). $A_i = 20$ ma $\div$ 5 ma; the current gain is 4. Note that this value agrees with that calculated by using the gain formula (par. 3–23*a*3, example 2). In that case the same tunnel diode was used with the same load resistor.



A. Signal source with internal resistance and load resistor



B. Tunnel diode added in parallel

FIG. 3–22. Circuits showing current and voltage designations for calculating gain formulas of parallel amplifier with signal source having internal resistance

3–25. Parallel Amplifier with Signal Source Resistance

In discussing parallel amplifiers (par. 3–23), it was stated that the voltage gain (A_v) is always one. This statement is true if the signal source has no internal resistance. With internal resistance there is current, voltage, and power gain (*a*, *b*, *c* below). Figure 3–22A shows a signal source (e_s) with internal resistance (R_s) and a load resistor (R_L). The current through R_L is designated i_1 and the voltage across R_L is designated e_1. Figure 3–22B is the equivalent circuit of the same signal source and load resistor with a

tunnel diode in parallel with the load resistor. In this case the current through and the voltage across R_L are designated i_2 and e_2, respectively.

a. Current Gain. The current gain (A_i) is defined as the ratio of current (i_2) through the load resistor paralleled by the tunnel diode, to the current (i_1) without the tunnel diode.

1. Write the expression for current gain:

$$A_i = \frac{i_2}{i_1}$$

2. The formula for current gain (derived in Appendix B) is:

$$A_i = \frac{R_D}{R_D - R_T}$$

Where R_T is the parallel resistance of R_s and R_L:

$$R_T = \frac{R_s R_L}{R_s + R_L}$$

3. A numerical example of the use of the formula (A_i) to calculate current gain follows:

Example. Given: $R_s = 20$ ohms
$R_L = 30$ ohms
$-R_D = 20$ ohms

Find A_i.

(*a*) Determine R_T (parallel resistance of R_s and R_L):

$$R_T = \frac{20 \times 30}{50}$$

$$R_T = 12 \text{ ohms}$$

(*b*) Substitute values in formula for A_i (3 above):

$$A_i = \frac{20}{20 - 12}$$

Therefore,

$$A_i = 2.5$$

b. Voltage Gain. Voltage gain (A_v) is defined as the ratio of voltages e_2 and e_1, as indicated in Fig. 3–22. The formula for voltage gain is derived in Appendix B and is found to be identical to the formula for current gain (*a* above); thus:

1. $$A_v = \frac{e_2}{e_1}$$

2. $$A_v = \frac{R_D}{R_D - R_T}$$

c. Power Gain. Power gain (G) is the product of current and voltage gain.

1. Write the formula for power gain:

$$G = A_i A_v$$

2. Substitute formulas (a and b above):

$$G = \frac{(R_D)^2}{(R_D - R_T)^2}$$

Note that power gain in this case equals the square of the current gain or the square of the voltage gain. Using the data calculated in $a3$ above, then

$$G = A_i^2$$

$$G = (2.5)^2$$

$$G = 6.25$$

SECTION V. OSCILLATION, PARALLEL ARRANGEMENT

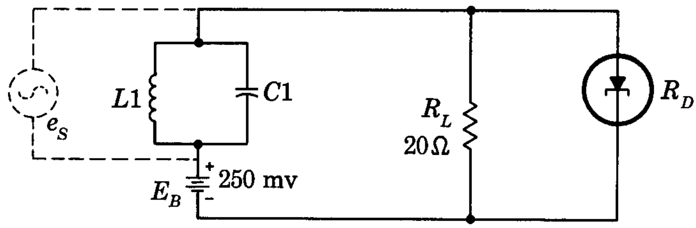
3–26. Parallel Oscillator

a. The formula for current gain in a parallel amplifier (par. 3–23) was determined to be as follows:

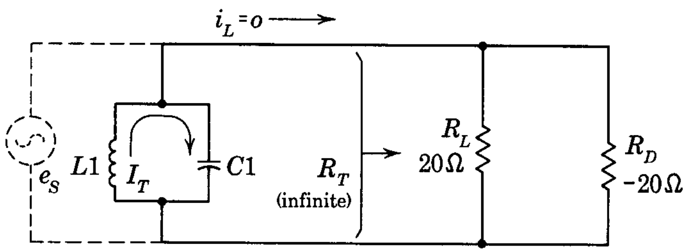
$$A_i = \frac{R_L}{R_L - R_D}$$

b. If load resistor (R_L) equals the negative resistance (R_D) of the tunnel diode, the denominator of the current-gain formula equals zero, indicating infinite gain. Oscillations occur instead.

c. Figure 3–23A shows a parallel oscillator. Considering the battery as an ac short, the equivalent circuit of the oscillator is shown in Fig. 3–23B. The tunnel diode being used has an ac negative resistance of 20 ohms (par. 3–11). The tank circuit (coil $L1$), the capacitor ($C1$), and the load resistor (R_L) are in parallel with the tunnel diode. Load resistor R_L represents the dc resistance of the tank circuit plus any loading of the tank circuit by coupling energy to an amplifier or antenna.



A. Parallel resonant oscillator



B. Equivalent circuit

FIG. 3–23. Parallel oscillator and its equivalent circuit

d. Note that the total resistance (R_T) *seen* by the tank circuit is infinite, calculated as follows:

$$
\begin{aligned}
R_T &= \frac{R_L(-R_D)}{R_L - R_D} \\
&= \frac{(20)(-20)}{20 - 20} \\
&= \frac{-400}{0} \\
&= \infty
\end{aligned}
$$

e. With infinite resistance, no ac current is drawn from the tank circuit. When the battery circuit (Fig. 3–23A) is closed, a tank current I_T is established and remains fixed because there are no losses in the tank circuit. The tank circuit acts like an ac voltage source (e_s) (shown in dashed lines) that delivers no current but varies the dc resistance of the tunnel diode at a desired time rate. This dc resistance variation *shifts* the battery voltage and power at the desired rate to make up for the power losses in resistor R_L (par. 3–27).

3–27. Parallel Oscillator, Power-Shift Relations

a. The reason for the start of oscillations in a parallel *amplifier* (par. 3–26) when the load resistance equals the negative resistance of the tunnel diode can be understood by studying the power shift relations.

b. Consider Fig. 3–11. Assuming the quiescent point marked × (15 ma, 250 mv), the maximum voltage swing is 200 mv [one half of 400 (450–50)]. The negative resistance (R_D) of the diode is 20 ohms (par. 3–11). The maximum power (P) that can be *shifted* by the tunnel diode can be determined as follows:

1. Write formula for power:

$$P = \frac{E^2}{R}$$

2. Substitute maximum voltage and value of R_D:

$$P = \frac{(0.2)^2}{20}$$

Note: 200 mv equals 0.2 volt. Therefore,

$$P = 2 \text{ mw}$$

c. The peak power (2 mw) that can be shifted will be compared to the power absorbed by using load resistors. Table 3–3 lists various load re-

TABLE 3–3

Load Resistor (ohms)	Power Absorbed (mw)
50	0.8
40	1
30	1.3
20	2
15	2.7
10	4
5	8

sistors, and the power absorbed if a voltage of 200 mv is across the load resistor.

d. Table 3–3 shows that a parallel load resistor of 5, 10, 15 ohms absorbs more power than the tunnel diode can *shift* (2 mw) at an ac rate from the battery. As a result oscillations cannot be sustained and will be quickly damped for lack of power. At 20 ohms (the same value as that of the negative resistance) and *higher* values, the tunnel diode can shift sufficient energy to sustain oscillations.

e. It can be stated therefore that any tunnel diode used in a parallel circuit will sustain oscillations if the parallel load resistor is *equal to or is greater than the negative resistance of the tunnel diode.* Because oscillations cannot be sustained at resistance values *less* than the negative resistance when in a parallel arrangement, the circuit is referred to as *short-circuit stable.*

SECTION VI. COMPOUND CONNECTED AMPLIFIERS

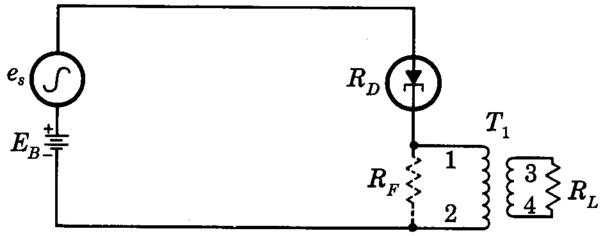
3–28. Impedance Matching

A study of the gain formulas for the series or parallel tunnel diode amplifiers (par. 3–12 and 3–23) indicates that maximum gain is achieved when the load resistor and the negative-resistance values are almost equal. (If equal, there will be oscillation.) A close approximation to impedance matching is therefore desirable. In many applications it will be necessary to use tunnel diodes and load resistors widely divergent in resistance values. Methods of bringing about impedance matching are by use of a transformer (*a* below), or by compound connections (*b* below). The advantages and disadvantages of each method are discussed in *c* below.

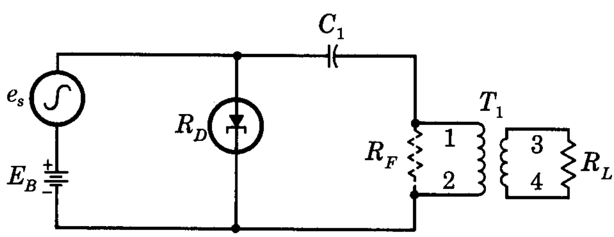
a. Transformer Matching. 1. Figure 3–24A shows a series amplifier using transformer T_1 to match the resistance of load resistor R_L to the negative resistance of the tunnel diode. Transformer T_1 can be a voltage step-up or step-down transformer, depending on whether the resistance of R_L is greater or less, respectively, than that of the diode. The main requirement is that the reflected resistance (R_F, shown in dashed lines) be slightly greater than the negative resistance of the diode. R_F must be slightly greater than R_D because the series amplifier is *open-circuit stable* (par. 3–16).

2. Figure 3–24B shows a parallel amplifier using transformer T_1 to match the resistance of load resistor R_L to the negative resistance of the tunnel diode. Capacitor C_1 is a dc blocking capacitor and prevents shorting of the tunnel diode bias by the low resistance of the transformer primary. The reflected resistance (R_F, shown in dashed lines) must be slightly *lower* than R_D because the parallel amplifier is *short-circuit stable* (par. 3–27).

b. Compound Connections. Compound connected series and parallel amplifiers are shown in Figs. 3–25 through 3–30 and discussed in detail in



A. Series amplifier, transformer impedance matching



B. Parallel amplifier, transformer impedance matching

FIG. 3–24. Series and parallel amplifiers using transformers to achieve impedance matching

paragraphs 3–29 through 3–33. In this method an impedance-matching resistor (R_Z) is used in parallel or series with the tunnel diode or the load to achieve virtual impedance matching and maximum gain. It will be seen that if the load resistor is much larger than the negative resistance of the tunnel diode, usually a series amplifier with an impedance-matching resistor will be used to achieve maximum gain (par. 3–29 and 3–30). If the load resistor is much smaller than the tunnel diode negative resistance, a parallel amplifier with an impedance-matching resistor most often will be used (par. 3–32 and 3–33).

c. Advantages and Disadvantages. 1. The main advantage of the transformer is that its power efficiency is very high. This fact is important in portable equipment where conservation of battery power is desirable. The transformer has a disadvantage, however, because it does not have as good a frequency response as resistors. The low-frequency gain is attenuated by the reduced inductive reactance of the windings; the high-frequency response is attenuated by leakage reactance between primary and secondary windings. Transformers are also more expensive and larger in size than resistors.

2. The main disadvantage of using resistors is the reduced power efficiency; this is especially important where batteries must supply the power. However, resistors provide advantages of high gain, good bandwidth, economy of parts, and small size.

3–29. Series Amplifier, Diode Paralleled

Figure 3–25A shows a series amplifier with impedance-matching resistor R_Z in parallel with the tunnel diode. Resistor R_L is the load resistor and capacitor $C1$ couples the output signal to the following stage. Note that R_L is 150 ohms and R_D is -100 ohms; this represents a large mismatch in impedance. Figure 3–25B is the equivalent circuit of the amplifier. Note that paralleling the diode with R_Z (302 ohms) *increases* the effective negative resistance in series with the load resistor. The combined resistance (R_{eq}) of R_Z and R_D in parallel is -149 ohms and is shown in dashed lines. The gain of the amplifier with and without R_Z is covered in *a* and *b* below, respectively. Only the voltage gain is discussed.

a. Gain Without Matching. To find the voltage gain (A_v) of this amplifier without R_Z in the circuit:

1. Write the voltage gain formula (par. 3–12*b*):

$$A_v = \frac{R_L}{R_L - R_D}$$

2. Substitute the given values:

$$A_v = \frac{150}{150 - 100}$$

Then

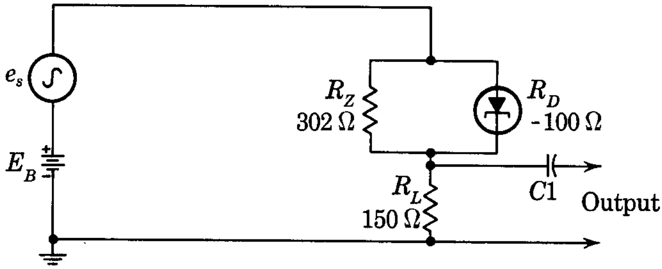
$$A_v = 3$$

b. Matched Gain. To find the gain of the amplifier with R_Z in the circuit, use the same formula (*a*1 above), but use 149 (R_{eq} for R_D):

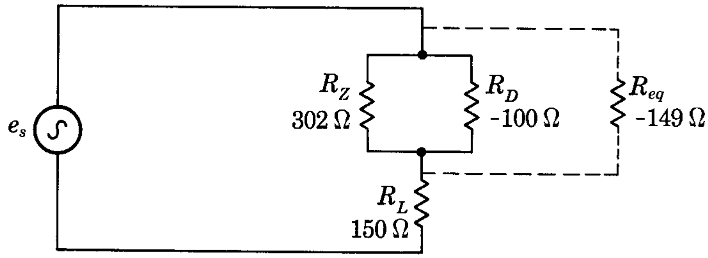
1.
$$A_v = \frac{150}{150 - 149}$$

$$A_v = 150$$

2. Note that the gain has increased 50 times ($150 \div 3$) by virtual impedance matching. The power efficiency has been reduced because of the current drain by resistor R_Z. However, the improved gain is usually more desirable.



A. Series amplifier, diode paralleled



B. Equivalent circuit

FIG. 3–25. Compound-connected series amplifier (diode paralleled), and its equivalent circuit

3–30. Series Amplifier, Load Paralleled

a. Figure 3–26A shows a series amplifier with impedance-matching resistor R_Z in parallel with load resistor R_L. As in the case of the same basic amplifier (without resistor R_Z) discussed in paragraph 3–29, the tunnel diode resistance (R_D) is -100 ohms and the load resistor is 150 ohms. The voltage gain of the basic amplifier (par. 3–29*a*2) is 3.

b. Figure 3–26B shows the equivalent circuit of the amplifier. Note that resistor R_Z (310 ohms) reduces the effective load resistance (R_{eq}) to 101 ohms. To find the gain (A_v), use the same formula (par. 3–12) but substitute R_{eq} for R_L thus:

1. Write the formula:

$$A_v = \frac{R_{eq}}{R_{eq} - R_D}$$

2. Substitute values:

$$A_v = \frac{101}{101 - 100}$$

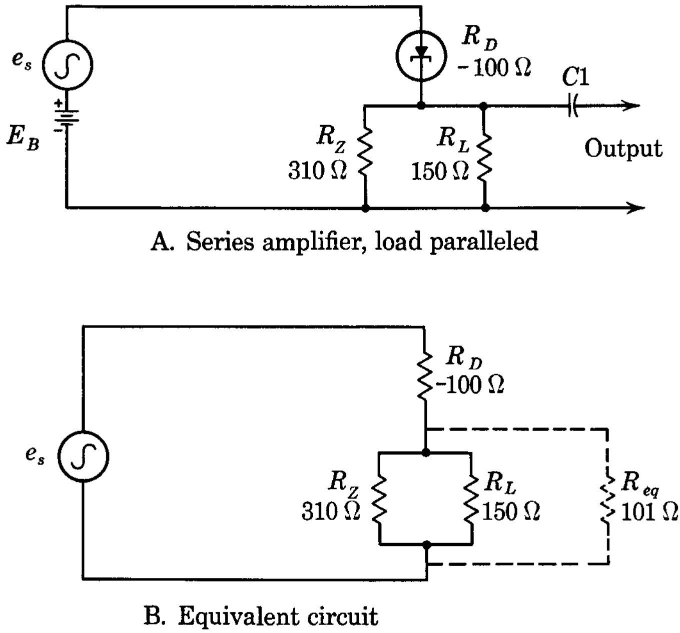


FIG. 3–26. Compound-connected series amplifier (load paralleled), and its equivalent circuit

Thus

$$A_v = 101$$

3. Note that the voltage gain has been *increased* approximately 33 times (101 ÷ 3) by using this method of impedance matching.

3–31. Series Amplifier, Series Matching Resistor

Figure 3–27A shows a series amplifier with impedance-matching resistor R_Z in series with load resistor R_L. Capacitor $C1$ is a dc blocking capacitor that couples the output signal to the following stage. Note that R_L is 75 ohms and R_D is -100 ohms. Not only does this condition represent a large mismatch that would reduce gain, but amplification would be impossible because the circuit would oscillate. The series amplifier is *open-circuit stable* (par. 3–16); the load resistor must be larger than the diode negative resistance to prevent oscillation. Voltage gain, therefore, without the matching resistor (R_Z, 26 ohms) cannot be calculated. Figure 3–27B shows the equivalent circuit of the amplifier. Note that the equivalent load (R_{eq}, shown in dashed lines) is 101 ohms. In *a* below the voltage gain across the equivalent load is calculated; in *b* below the actual voltage gain (across load resistor R_L) is calculated.

a. Gain Across Equivalent Load. To find the voltage gain across the equivalent load (R_{eq}):

1. Write the voltage gain formula (par. 3–12*b*), but substitute R_{eq} for R_L:

$$A_v = \frac{R_{eq}}{R_{eq} - R_D}$$

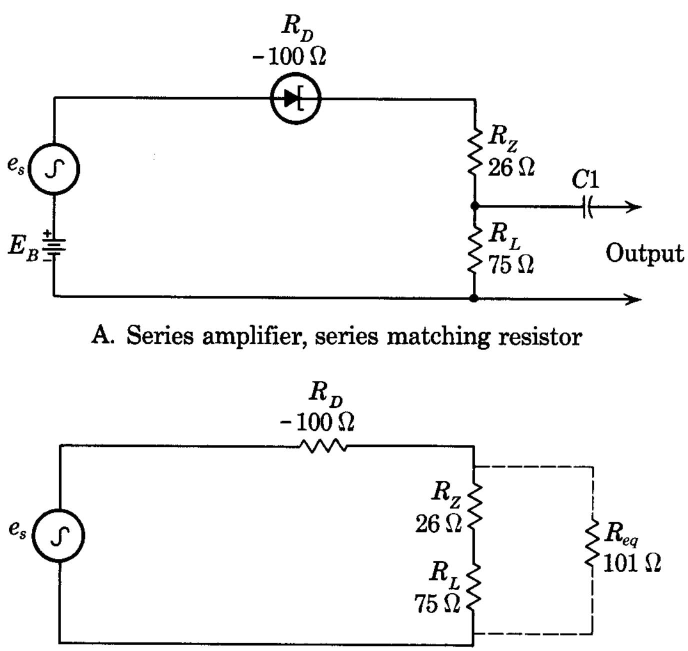
2. Substitute the given values:

$$A_v = \frac{101}{101 - 100}$$

Then

$$A_v = 101$$

b. Gain Across Load (R_L). The voltage gain (A_{vL}) across the load (R_L) is the ratio of the load resistance (R_L) to the equivalent resistance (R_{eq}) multiplied by the gain across R_{eq} (*a* above).



A. Series amplifier, series matching resistor

B. Equivalent circuit

FIG. 3–27. Series amplifier (with matching resistor in series with load), and its equivalent circuit

1. Express A_{vL} as a formula:

$$A_{vL} = \frac{R_L}{R_{eq}} \times A_v$$

2. Substitute known values:

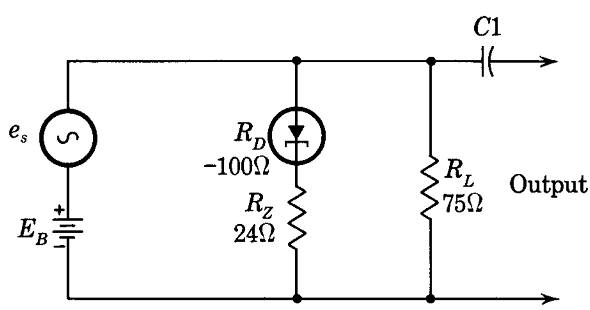
$$A_{vL} = \frac{75}{101} \times 101$$

Then

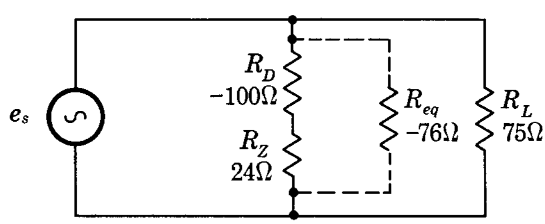
$$A_{vL} = 75$$

3–32. Parallel Amplifier, Diode Series Resistor

Figure 3–28A shows a parallel amplifier with impedance matching resistor (R_Z) in series with the tunnel diode. Resistor R_L is the load resistor, and capacitor $C1$ couples the signal to the following stage. Note that R_L is 75 ohms and R_D is -100 ohms. Figure 3–28B is the equivalent circuit of the amplifier. Placing R_Z (24 ohms) in series with the tunnel diode *decreases*



A. Parallel amplifier, diode series resistor



B. Equivalent circuit

FIG. 3–28. Compound-connected parallel amplifier (resistor in series with diode), and its equivalent circuit

the effective negative resistance (R_{eq}) in parallel with R_L to 76 ohms (100 − 24). Resistance R_{eq} is shown in dashed lines. The *current* gain of the amplifier with and without R_Z is covered in *a* and *b* below, respectively.

a. Gain Without Matching. To find the gain (A_i) of the amplifier without R_Z:

1. Write the current gain formula (par. 3–23):

$$A_i = \frac{R_D}{R_D - R_L}$$

2. Substitute values:

$$A_i = \frac{100}{100 - 75}$$

Then

$$A_i = 4$$

b. Matched Gain. To find the gain of the amplifier with R_Z in the circuit, use the same formula (*a*1 above), but use 76 ohms (R_{eq}) for R_D:

1. Write the formula:

$$A_i = \frac{R_{eq}}{R_{eq} - R_L}$$

2. Substitute values:

$$A_i = \frac{76}{76 - 75}$$

Then

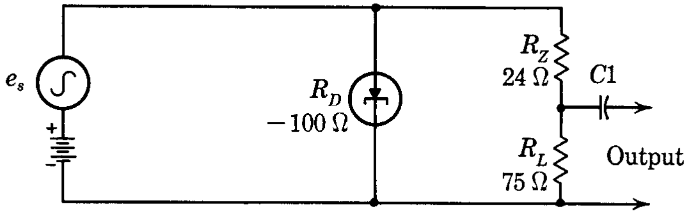
$$A_i = 76$$

Note that the gain has increased 19 times (76 ÷ 4) with R_Z in series with the tunnel diode.

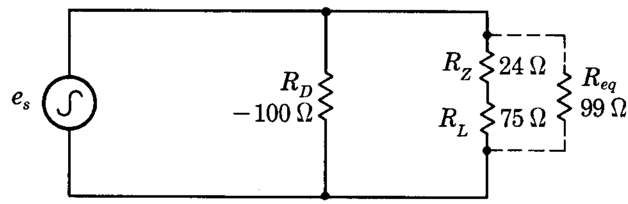
3–33. Parallel Amplifier, Load Series Resistor

a. Figure 3–29A shows a parallel amplifier with impedance-matching resistor R_Z in series with load resistor R_L. As in the case of the same basic amplifier (without resistor R_Z) discussed in paragraph 3–32, the tunnel diode resistance (R_D) is −100 ohms and the load resistor R_L is 75 ohms. The current gain of the basic amplifier is 4 (par. 3–32*a*2).

b. Figure 3–29B shows the equivalent circuit of the amplifier. Note that resistor R_Z (24 ohms) *increases* the effective load resistance (R_{eq}) to 99 ohms. To find the gain (A_i) use the same formula (par. 3–23) but substitute R_{eq} for R_L thus:



A. Parallel amplifier, resistor in series with load



B. Equivalent circuit

FIG. 3–29. Compound-connected parallel amplifier (resistor in series with load), and its equivalent circuit

1. Write the formula:

$$A_i = \frac{R_D}{R_D - R_{eq}}$$

2. Substitute values:

$$A_i = \frac{100}{100 - 99}$$

Then

$$A_i = 100$$

Note that the current gain has been increased 25 times (100 ÷ 4).

3–34. Parallel Amplifier, Parallel Matching Resistor

Figure 3–30A shows a parallel amplifier with impedance-matching resistor R_Z in parallel with load resistor R_L. Capacitor $C1$ is a dc blocking capacitor that couples the output signal to the following stage. Note that R_L is 150 ohms and R_D is -100 ohms. Not only is this condition a large mismatch, but amplification would be impossible because the circuit would oscillate. The parallel amplifier is *short-circuit stable* (par. 3–27); the load resistor must be smaller than the diode negative resistance to prevent oscil-

lation. Current gain, therefore, without the matching resistor (R_Z, 291 ohms) cannot be calculated. Figure 3–30B shows the equivalent circuit of the amplifier. Note that the equivalent load (R_{eq}, shown in dashed lines) is 99 ohms. In *a* below the current gain through the equivalent load is calculated; in *b* below the actual current gain (through load resistor R_L) is calculated.

a. Gain Through Equivalent Load. To find the current gain (A_i) through the equivalent load (R_{eq}):

1. Write the current gain formula (par. 3–23), but substitute R_{eq} for R_L:

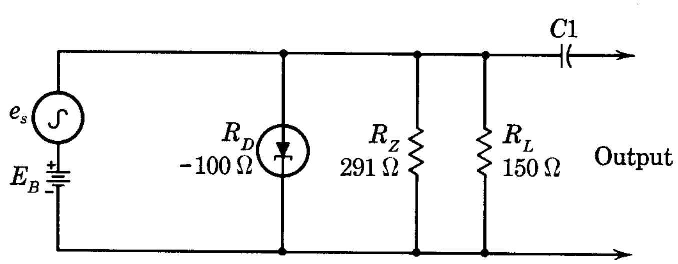
$$A_i = \frac{R_D}{R_D - R_{eq}}$$

2. Substitute the given values:

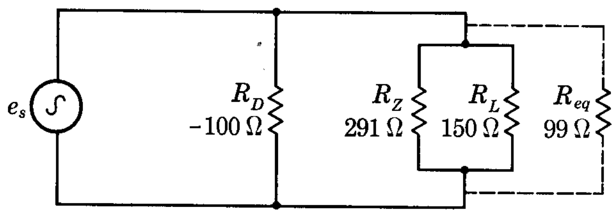
$$A_i = \frac{100}{100 - 99}$$

Then

$$A_i = 100$$



A. Parallel amplifier with parallel matching resistor



B. Equivalent circuit

FIG. 3–30. Parallel amplifier (with parallel matching resistor) and its equivalent circuit

b. Gain Through Load (R_L). The current gain (A_{iL}) through the load resistor is the ratio of the load *conductance* ($1/R_L$) to the equivalent load conductance ($1/R_{eq}$) multiplied by the gain through R_{eq} (*a* above).

1. Express A_{iL} as a formula:

$$A_{iL} = \frac{1/R_L}{1/R_{eq}} \times A_i$$

2. Substitute known values:

$$A_{iL} = \frac{1/150}{1/99} \times 100$$

$$A_{iL} = \frac{99}{150} \times 100$$

Then

$$A_{iL} = 66$$

3–35. Summary

a. Circuit arrangements exhibiting ac negative resistance were investigated and applied as early as 1918.

b. The transitron oscillator is an example of a pentode electron vacuum tube arranged in a circuit to exhibit negative resistance.

c. Examples of semiconductor devices other than the tunnel diode that exhibit ac negative resistance are the point-contact transistor, the four-layer diode, the *p-n-p-n* transistor, and the unijunction transistor.

d. Ac negative resistance is a second-order (mathematical) concept. It represents mathematically (for computation purposes) the ac *effect* in a given equation of a varying dc resistance. The varying dc resistance is such that increasing voltage results in decreasing current; i.e., increased dc voltage results in increased dc resistance.

e. On a mathematical (second order) basis a dc resistance which varies as indicated (*d* above) *acts like* a voltage or current generator, or an ac negative resistance.

f. A series amplifier consists of a bias battery, a signal source, and a tunnel diode in series with a load resistor. The signal source voltage varies the dc resistance of the tunnel diode so as to *shift* battery power to the load resistor; i.e., an amplified voltage-signal appears across the load resistor.

g. If the series amplifier is compared to the same circuit without the tunnel diode, there is current, voltage, and power gain (par. 3–12). If the input signal from the signal source and the output signal across the load resistor only are considered, then a series amplifier has only voltage and power gain (par. 3–14); the current gain is one (Fig. 3–21).

h. By using the negative-resistance concept in the equivalent circuit of a series amplifier (par. 3–12), desired gain formulas can be derived.

i. A series amplifier will oscillate if the load resistance is equal to or is less than the tunnel diode negative resistance. If the signal source is replaced by a tank circuit, an oscillator is formed.

j. The series amplifier is open-circuit stable or short-circuit unstable.

k. When a properly biased tunnel diode is ac short-circuited, it may oscillate. The unloaded tunnel diode usually oscillates at 1 to 10 kmc, depending on the parameters of its equivalent circuit (Fig. 3–15).

l. The self-resonant frequency (f_s) of a tunnel diode is the frequency at which the reactance of its equivalent circuit (Fig. 3–12) equals zero. The resistance cutoff frequency (f_r) of a tunnel diode is the frequency at which more power is dissipated by the circuit than can be *shifted* by the negative resistance from the dc source.

m. A tunnel diode parallel amplifier displays current and power gain. The voltage gain is one. If the signal source has internal resistance, then the voltage gain is greater than one.

n. By using the ac negative-resistance concept in an equivalent circuit, desired gain formulas for the parallel amplifier may be derived.

o. If the load resistance of a tunnel diode parallel amplifier equals, or is *greater* than, the ac negative resistance of the tunnel diode, oscillations will occur. If the signal source is replaced by a tank circuit, an oscillator is formed.

p. The tunnel diode parallel amplifier is short-circuit stable or open-circuit unstable.

q. Maximum gain in a series amplifier will be achieved if the load resistance is only slightly greater than the ac negative resistance (virtual impedance match). Maximum gain in a parallel amplifier will be achieved if the load resistor is only slightly less than the ac negative resistance (virtual impedance match).

r. Virtual impedance match can be achieved in tunnel diode amplifiers by use of transformers or compound (resistor diode) connections.

Chapter 4

AMPLIFIER CONSIDERATIONS AND CIRCUITS

SECTION I. NOISE CONSIDERATIONS

4–1. General

a. This chapter covers several of the most important aspects of the tunnel diode as an amplifying device:

1. Noise considerations (par. 4–2 and 4–3).
2. Distortion (par. 4–4, 4–5, and 4–6).
3. Temperature characteristics (par. 4–7, 4–8, and 4–9).
4. Bias requirements (par. 4–10, 4–11, and 4–12).

b. Several practical tunnel diode amplifiers are covered in paragraphs 4–13 through 4–16.

4–2. Origin of Tunnel Diode Noise

Noise refers to an irregular electrical signal (having voltage, current, and power) to which no specific frequency can be assigned. Converted to sound by a speaker, noise manifests itself as a sizzling, crackling, or frying sound. On a radar scope, it is referred to as *grass.* On a television screen it is referred to as *snow.* The noise power present in an amplifier is important because the noise power present determines the minimum signal power-handling capacity of the amplifier. The smaller the noise power, the smaller the signal power necessary to override the noise. Studies of the noise characteristics of the tunnel diode show that the tunnel diode, like the electron tube, exhibits both shot noise (*a* below) and thermal noise (*b* below).

a. Shot Noise. In an electron tube, shot noise is the result of the irregular impact of electrons on the plate of the tube, the electrons acting very much like hailstones falling on a pavement. In the tunnel diode, shot noise is manifested during the tunneling process. The shot noise can be considered

the result of the impact of the carriers (electrons and holes) on the barrier (Ch. 2). It is believed that shot noise is particularly the result of carriers (electrons and holes) deflected from the barrier when unlike charges of equal energy levels are not on opposite sides of the barrier. Figure 4–1 shows the current-voltage curve of a tunnel diode and several curves showing the approximate noise figure (par. 4–3) of a typical tunnel diode taken at the indicated frequencies. The noise figure depends on the frequency being amplified. In each case, however, note that the noise figure rises at the *valley current* of the tunnel diode. At the valley current there is also the maximum deflection of carriers from the barrier; see Figs. 2–22 and 2–23.

b. Thermal Noise. In a conductor or resistor, the free electrons are continuously in motion at a velocity that depends upon the temperature. At a given moment a larger number of electrons will be moving in one direction than the number moving in the other direction. This motion causes a noise voltage to be developed at the end of the conductor or the resistor which corresponds to the direction of major electron flow. The voltage will vary instantaneously and irregularly to produce the characteristic noise voltage. The square of the noise voltage is directly proportional to the resistance of the conductor or the resistor. In paragraph 3–17 the equivalent circuit of the tunnel diode was studied; it was shown that a parasitic resistance (R_s) is associated with the tunnel diode leads. This resistance and the resistance of the bulk material of the tunnel diode contribute to thermal noise in the tunnel diode (par. 4–3).

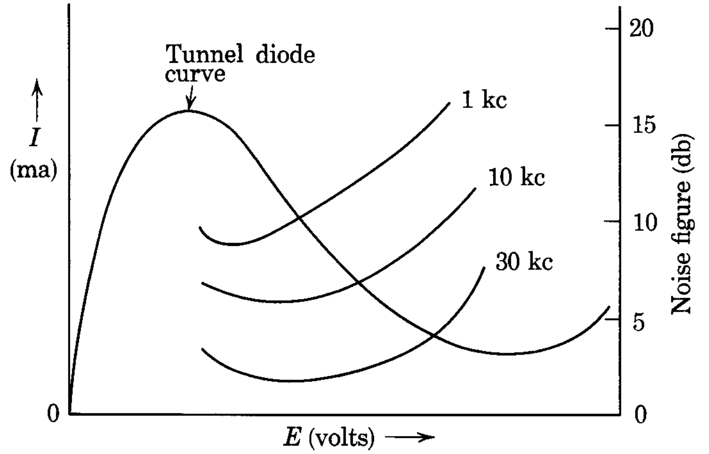


FIG. 4–1. Tunnel diode current-voltage curve and noise figure curves at low frequencies

4–3. Tunnel Diode Noise Figure

The noise figure of merit (F) of the tunnel diode evaluates the noise producing characteristics of the tunnel diode alone. The formula for F which permits a quantitative calculation of the value of the noise figure is presented and discussed in detail in Appendix B. This formula relates the effects of shot noise caused by tunneling, thermal noise caused by the parasitic dc resistance of the diode, and the particular operating frequency as related to the resistance cutoff frequency (f_r) of the diode (par. 3–17). Study of this formula leads to the following conclusions:

a. In and immediately above audio frequencies the noise is due almost entirely to shot noise and is inversely proportional to the operating frequency (Fig. 4–1). Below 30 kc the noise figure is relatively large. At 30 kc a low average noise figure of 3 to 4 db is obtained.

b. At operating frequencies below half the resistance cutoff frequency (f_r), the noise characteristics are due mainly to thermal noise and satisfactory values of $F = 3$ db are obtained. The higher the frequency the more pronounced the thermal noise because of the shunting effect of the inherent capacitance (C_D) of the diode (par. 3–17) on the ac negative-resistance parameter (R_D).

c. When the operating frequency (f) reaches $0.707f_r$, the noise figure of merit reaches 6 db. When f equals $0.9f_r$, the noise figure reaches 10 db. For low noise operation the tunnel diode performs best at operating frequencies above 25 kc and below $0.707f_r$.

SECTION II. DISTORTION

4–4. General

a. In a three-terminal device, such as an electron tube or a transistor, the amount of distortion experienced by a signal amplified by the device can be determined by studying the *dynamic transfer characteristic curve* of the device in relation to the signal introduced. In the case of the electron tube, this curve plots the plate current against the control grid voltage under load conditions. In the case of the transistor (Appendix C), this curve plots the collector current against the base current under load conditions. In each case operation over the most linear portion of the dynamic transfer characteristic curve results in minimum distortion. Operation over the nonlinear portion of the dynamic transfer characteristic curve will result in distortion. The amount of distortion is generally indicated by projecting the input signal onto the dynamic transfer characteristic curve and, from the projection, drawing the output wave shape. This procedure is covered in many textbooks on electron tubes and transistors.

b. Because the tunnel diode is a two-terminal device, a dynamic transfer characteristic curve in the sense employed in *a* above is not possible. In this case, distortion or nondistortion is revealed by studying the ac negative-resistance curve of the tunnel diode as derived from its current-voltage curve. In paragraph 4–5 an idealized (all straight-line portions) current-voltage chart is considered; the resistance curve is derived and discussed with respect to distortion. In paragraph 4–6 the current-voltage chart of a practical germanium tunnel diode is considered; the resistance curve is derived and also discussed.

4–5. Ideal Linear Characteristic

a. Figure 4–2A shows the idealized current-voltage curve of the tunnel diode. Each portion of the curve is a straight line, representing a constant ratio of voltage to current for the length of the straight-line portion. The resistance therefore for each portion is a constant value. Figure 4–2B shows the corresponding resistance values plotted against the same horizontal voltage axis as that of the current-voltage chart. From zero current to peak current (I_p) a constant resistance of 55 ohms is displayed. At the I_p point the resistance changes abruptly to a negative value of 298 ohms and remains constant to the valley current (I_v) point. Note that this represents the ac negative resistance and is calculated by using incremental dc values (par. 3–10). At the I_v point the resistance changes abruptly to a constant positive resistance.

b. Operation of the tunnel diode as an amplifier is confined to the negative-resistance portion of the curve. Large and small input signals are shown projected onto the negative-resistance line. In this case, no distortion will be introduced for the small or the large signal because the resistance is constant over the complete length of the voltage signal.

4–6. Nonlinearity and Distortion

a. The actual current-voltage curve of a germanium tunnel diode is shown in Fig. 4–3A. The corresponding ac resistance is plotted in Fig. 4–3B. As in the previous case (par. 4–5), the resistance is positive from zero to peak current, negative from peak to valley current, and positive thereafter. However, the resistance does not remain constant over any portion of the curve. The resistance at any point is calculated by using a small right-angle triangle of incremental values of current and voltage. Several small triangles are shown on the zero to I_p portion of the curve. Note that as the peak is approached, the incremental value of current approaches zero and that of the voltage remains constant; the resistance value ($\Delta E/\Delta I$) approaches infinity and reaches infinity when the peak is reached [$(\Delta E/0) = \infty$)] Immediately after the peak the resistance is extremely high and

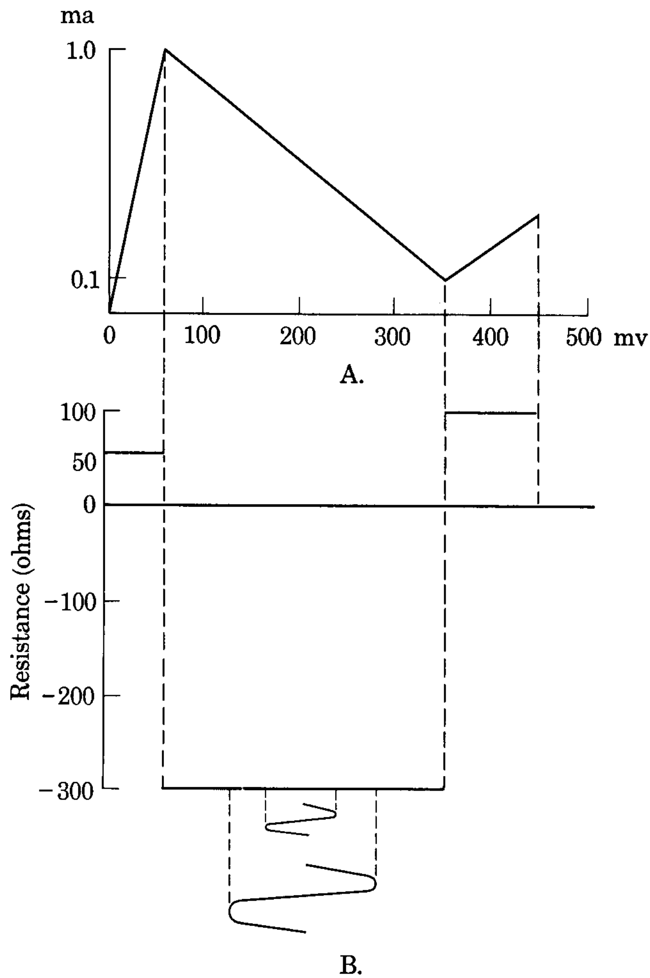


FIG. 4–2. Idealized current-voltage curve and corresponding ac resistance

negative. It remains negative until the valley point (I_v) is reached. At this point the resistance goes through infinity from a high negative value to a high positive value.

b. The curvature of the negative-resistance portion of the curve (Fig. 4–3B) shows that the negative resistance starts at a high value, decreases to a minimum value, and then increases again to a high value. The point at which the negative resistance reaches a minimum point (marked by a dot) is referred to as the *inflection* (turning) point. The corresponding inflection point on the current-voltage curve is also marked by a dot.

The inflection point is centered at the most linear portion of the curve. Most amplifiers are biased for operation at this point.

c. For a small signal input as shown, the distortion is minimum and the negative resistance can be considered to be constant and equal to the value (−151 ohms) at the inflection point. For a larger signal input the distortion is larger and the negative resistance must be considered the average of the values over the operating portion; in this case the average negative resistance is −285 ohms. The greater the difference in the negative resistance at the inflection point and the average negative resistance over the operating range, the larger the distortion.

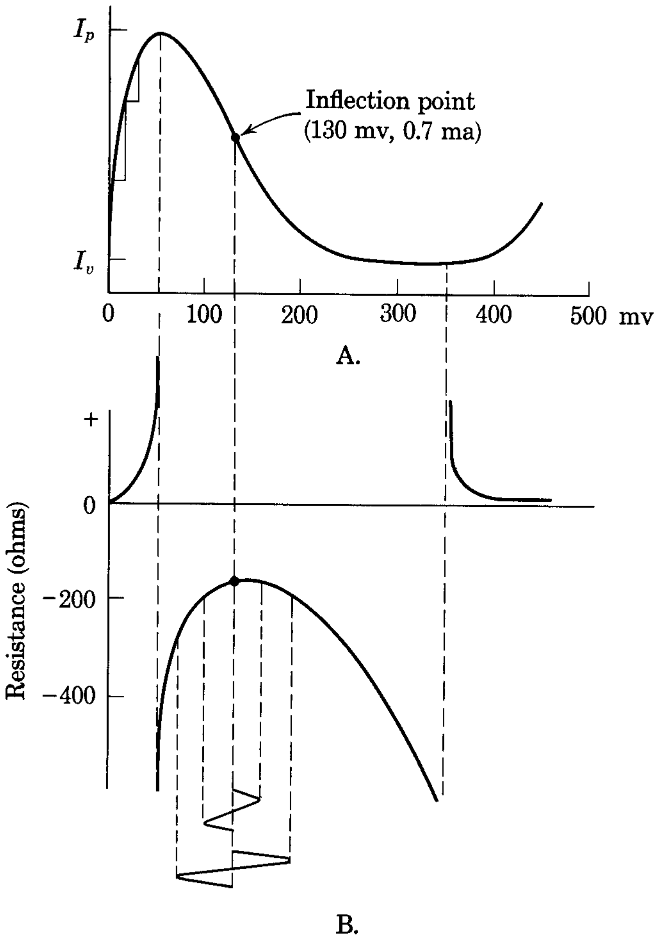


FIG. 4-3. Germanium tunnel diode current-voltage curve and corresponding ac resistance

SECTION III. TEMPERATURE CHARACTERISTICS

4–7. Temperature Effects on Peak and Valley Currents

a. Peak Current. The peak current (I_p) of a tunnel diode may increase or decrease with increasing temperature. Whether the peak current increases or decreases depends upon the types of impurities used to produce the *n*- and *p*-type semiconductors of which the diode is made. Most tunnel diodes available for commercial use display the peak-current and temperature characteristics shown in the upper portion of Fig. 4–4. In this case maximum peak current occurs at room temperature (25° C) and decreases above and below this temperature. The effect of variation of peak current with temperature in applications such as amplifiers, oscillators, and switching circuits is discussed in paragraph 4–9.

b. Valley Current. Unlike the peak current whose temperature characteristics may rise or fall with temperature, the valley current increases with temperature. This characteristic is shown in the lower portion of Fig. 4–4. For germanium tunnel diodes, the rate of rise of valley current is approximately 0.75% per °C. Its effect in particular applications is discussed in paragraph 4–9.

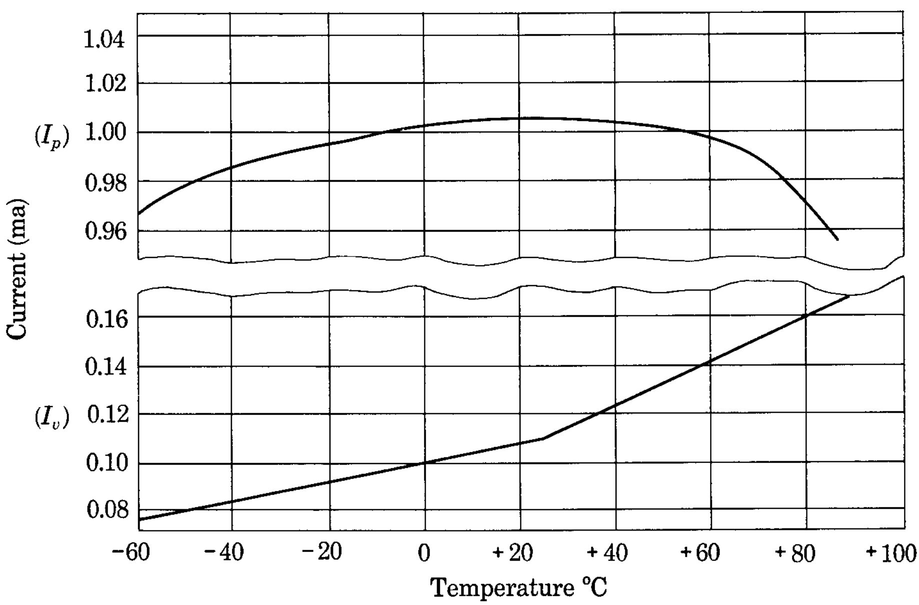


FIG. 4–4. Valley current and (typical) peak-current temperature characteristics

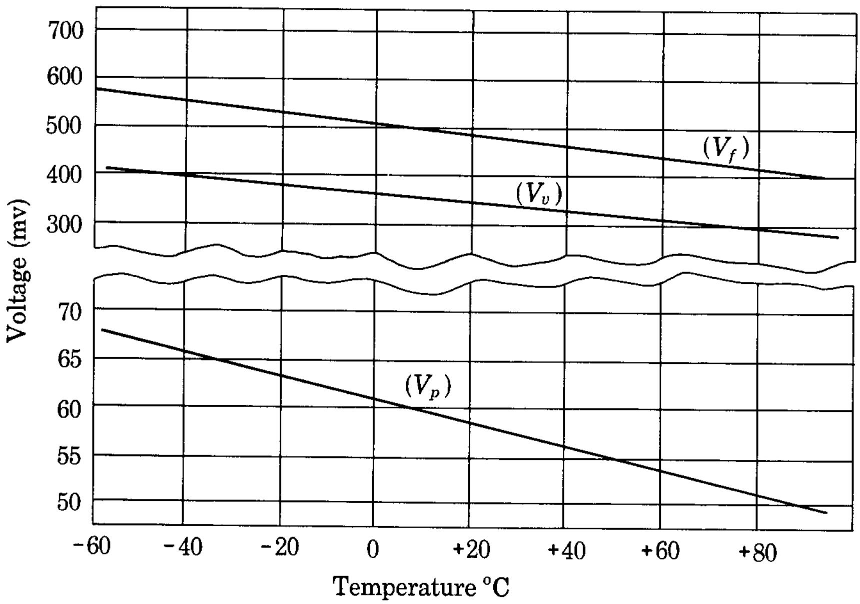


FIG. 4–5. Voltage-temperature characteristics of a typical germanium tunnel diode

4–8. Temperature Effects on Diode Voltages

The peak voltage (V_p), valley voltage (V_v), and forward voltage (V_f) temperature characteristic curves of a typical germanium tunnel diode are shown in Fig. 4–5. In each case the voltage decreases with increasing temperature. The effects of these variations in particular applications are discussed in paragraph 4–9. The typical temperature coefficient of each voltage is as follows:

a. Peak voltage (V_p) decreases 80 μv per °C of increasing voltage.
b. Valley voltage (V_v) decreases 0.9 mv per °C of increasing temperature.
c. Forward voltage (V_f) decreases 1 mv per °C of increasing temperature.

4–9. Temperature Effects in Specific Applications

The variations in tunnel diode parameters (par. 4–7 and 4–8) cause the current-voltage curve of the tunnel diode to vary in contour with temperature. The current-voltage curve of a typical germanium tunnel diode is shown in Fig. 4–6 as it appears at −50° C, +25° C, and +100° C. Note that the peak current is maximum at room temperature (25° C) and decreases above and below this temperature; the valley current rises with increasing temperature; and the peak, valley, and forward voltages decrease

with rising temperature. The importance of these effects in specific applications is discussed in *a*, *b*, and *c* below.

a. Amplifiers. 1. Tunnel diode amplifiers are usually biased at the inflection point (par. 4–6). In Fig. 4–6 the inflection point is marked by a dot on each current-voltage curve. Note that the voltage at the inflection point varies with temperature. A bias that is fixed at the inflection point at one temperature will not be at the inflection point at a different temperature. Extreme temperature variations in amplifiers with fixed bias will result in distortion. This problem can be overcome by using a bias source (par. 4–11) that varies with temperature to compensate for the voltage shift of the inflection point.

2. It was shown in Chapter 3 that maximum gain in a tunnel diode amplifier occurs when there is a virtual impedance match between the load resistance and the tunnel diode negative resistance. The temperature variations in the tunnel diode current-voltage chart cause an increase in negative resistance of 0.5% per °C. A relatively constant impedance match can be maintained by varying the bias to compensate, by using thermistors as load or matching resistors, or by employing negative feedback.

b. Oscillators. Impedance matching in oscillators is not required to sustain oscillations. Oscillations will be sustained (refer to Ch. 2) as long as the ac energy absorbed by the positive resistance of the circuit is not greater than the energy that can be shifted from the dc source by the negative resistance of the diode. Since the tunnel diode negative resistance decreases with temperature (*a*2 above), the energy that it can shift decreases

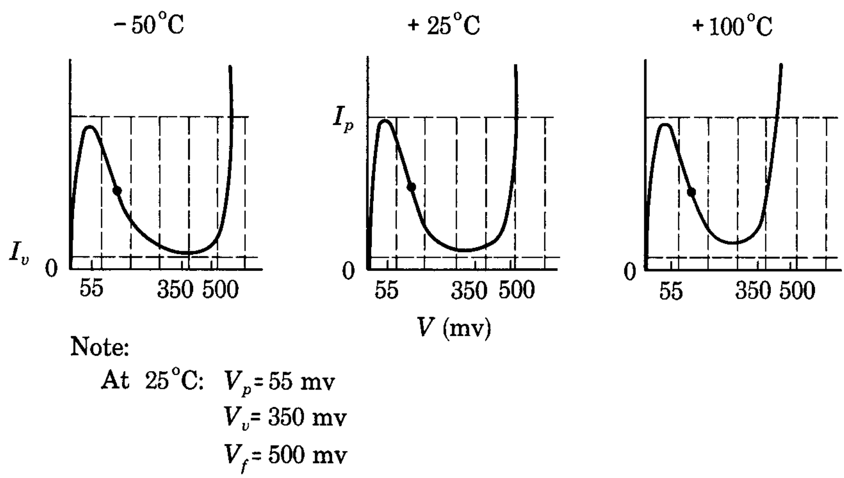


FIG. 4–6. Tunnel diode current-voltage curve at −50° C, +25° C, and +100° C

at the lower temperatures. In this case it is necessary only to ensure that at the lowest temperature of operation of the oscillator, sufficient energy can be shifted by the tunnel diode from the dc source to compensate for losses in the resistance of the circuit. Tunnel diode oscillators have been operated successfully from 4° K (Kelvin) to 573° K.

c. Switching Circuits. The primary requirement of switching circuits (Ch. 8) is the stability of the peak current (I_p). Switching circuits are normally biased below the peak current point for the *on* condition. *Turning off* (or switching) the circuit is determined by an input pulse having a fixed current amplitude. This current pulse raises the total current through the tunnel diode to the peak current, an unstable point, which causes the circuit to switch to the *off* condition, located in the forward voltage region. Variations in the peak current of the diode, then, will cause switching at different levels of input pulse current. Variations in the forward voltage will also cause variations in the output of the switching circuit. For switching circuit applications, tunnel diodes must be selected that have maximum peak-current temperature stability. In addition, temperature compensating devices can be employed to minimize output voltage variations.

SECTION IV. BIAS CONSIDERATIONS

4–10. General

a. For operation as an amplifier, the tunnel diode is usually biased at the inflection point. This point is at the center of the more linear portion of the current-voltage chart [the point of minimum negative resistance (Fig. 4–3)]. At this point the maximum voltage swing and, therefore, the maximum dynamic range are obtained. For germanium tunnel diodes the inflection point occurs between 125 and 140 mv; the more linear portion is between 80 and 180 mv, providing a voltage swing of 100 mv peak-to-peak.

b. The gain of a tunnel diode amplifier as discussed in Chapter 3 is primarily dependent on the negative resistance of the diode. A study of Fig. 4–3 shows that slight variations in bias voltage can cause large variations in the negative resistance and, therefore, in the gain of an amplifier. To avoid large variations in gain, it is necessary to bias the tunnel diode amplifier from a *stable* voltage source. If batteries are to be used, mercury cells are ideal because they supply a fixed voltage over a long operating time due to the fact that little internal resistance is developed with use. Otherwise stable voltages can be obtained by:

1. Using Zener diodes as voltage regulators (par. 4–11).
2. Using *forward-biased* rectifying diodes as voltage regulators (par. 4–12).

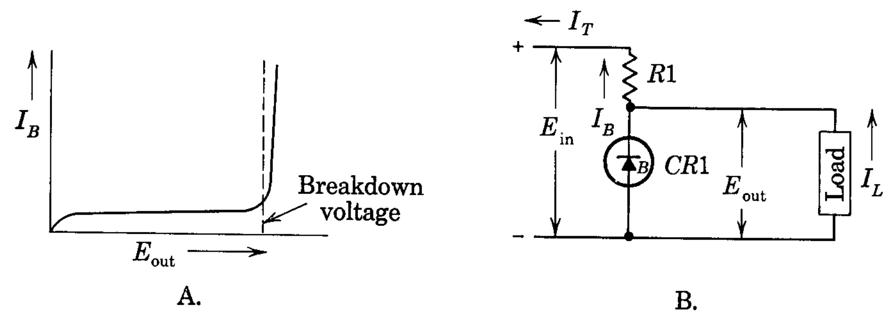


FIG. 4–7. Breakdown diode current-voltage chart and voltage regulator

4–11. Breakdown Diode Voltage Regulator

a. The current through and the voltage across a reverse-biased junction diode is shown in Fig. 4–7A. At a certain value of reverse-bias voltage (E_{out}), the current increases rapidly while the voltage across the diode remains almost constant. This voltage is called the breakdown, or Zener, voltage, and the diode is called a *breakdown* or *Zener* diode. This behavior is similar to that of some gas-filled tubes that start conducting at a particular voltage and continue to conduct varying amounts of current while the voltage across the tube remains almost fixed. As in the case of the gas-filled tube, the breakdown diode can be used as a voltage regulator (Fig. 4–7B). Variations in input current (I_T) or load current (I_L) are compensated by variations in breakdown diode current (I_B). This action maintains an almost fixed output voltage (E_{out}). Depending on the particular diode, the breakdown voltage can be any value from 2 to 100 volts.

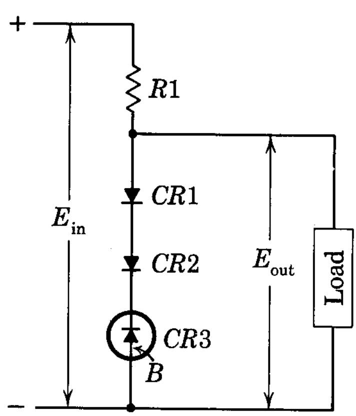


FIG. 4–8. Breakdown diode temperature-compensated voltage regulator

b. The reverse-biased junction diode has a negative temperature coefficient of resistance provided that the reverse-bias voltage does not equal or exceed the breakdown voltage. The breakdown diode has a *positive* temperature coefficient of resistance several times larger than the *negative* temperature coefficient of resistance of the forward- or reverse-biased junction diode. To compensate for the positive temperature coefficient of the breakdown diode, forward-biased diodes can be placed in series with the breakdown diode. Figure 4–8 shows a circuit with

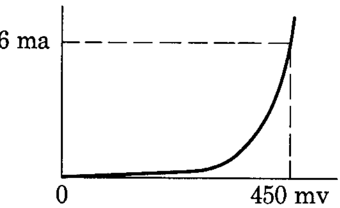
two forward-biased diodes ($CR1$ and $CR2$) in series with the breakdown diode ($CR3$). The total resistance of the three diodes remains constant over a wide range of temperatures. The complete result is a constant voltage output, although temperature, input voltage (E_{in}), and load current drain may vary.

4–12. Forward-Biased Diode Regulator

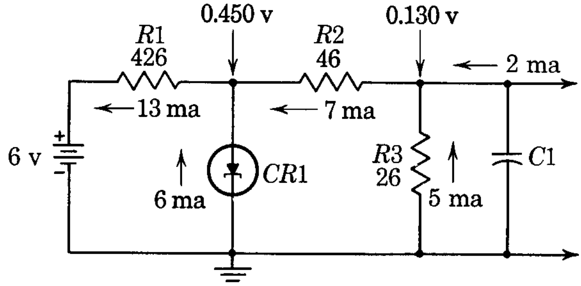
a. A forward-biased rectifying diode (Fig. 4–9) can be used as a voltage regulator. As in the case of the Zener diode, a large variation in current can occur in the forward direction while the voltage drop remains relatively constant (Fig. 4–9A). Again this characteristic is similar to that of a gas-tube regulator. The forward-biased diode is more desirable as a voltage regulator for tunnel diodes than the Zener diode for two reasons; first, the voltage drop is lower and closer to the biasing voltage of the tunnel diode; and second, its temperature characteristic helps compensate for the temperature characteristic of the tunnel diode (*c* below).

b. Figure 4–9B shows a forward-biased germanium diode used in a voltage regulator circuit. The diode is biased at 450 mv and the output voltage is 130 mv. Resistor $R1$ drops the battery voltage to the forward bias of the germanium diode. Resistors $R2$ and $R3$ form a low-impedance voltage divider to drop the output voltage to the desired value. The value of bypass capacitor $C1$ depends upon the operating frequency of the amplifiers being biased.

c. Because the forward-biased diode has a negative temperature coefficient of resistance, the voltage across the diode drops at the rate of 2.5 mv per °C of increasing temperature. In the tunnel diode the negative resistance increases 0.5% per °C. If the operating point is selected above the inflection point (about 145 mv), the increase in negative resistance with increasing temperature will be compensated for by a *decreasing* bias that



A.



B.

FIG. 4–9. Forward-biased germanium rectifying diode used as a voltage regulator

moves the operating point on the current-voltage chart to a *decreased* (original value of) negative resistance. The negative resistance therefore remains constant over a wide temperature range. If the decrease of 2.5 mv per °C of the forward-biased diode is too large a compensation, a germanium tunnel diode can be used as a voltage regulator. The voltage of the tunnel diode drops 1 mv per °C of increasing temperature.

SECTION V. AMPLIFIERS

4-13. Tuned RF Amplifier

a. A tuned RF amplifier stage is shown in Fig. 4-10A. Selectivity is provided by the parallel resonant circuit consisting of capacitor $C3$ and coil L1. The signal source consists of voltage generator e_s which has an internal resistance R_s. Note that this signal source could represent the out-

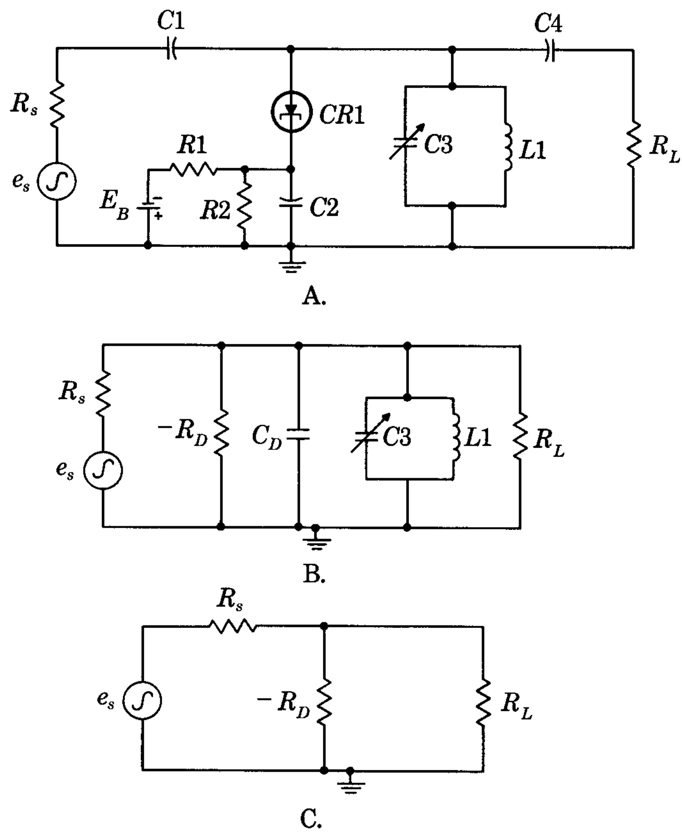


FIG. 4-10. RF tuned tunnel diode amplifier and its equivalent circuits

put of a transistor amplifier stage, or an antenna circuit. An amplified signal appears across parallel load resistor R_L because of the presence of tunnel diode $CR1$. Capacitors $C1$ and $C4$ are dc blocking capacitors. Resistors $R1$ and $R2$ form a voltage divider for biasing tunnel diode $CR1$. Capacitor $C2$ ac shorts resistor $R2$ to ground.

b. Figure 4–10B shows the same amplifier with the dc components removed. The components that have negligible effect on the ac signal are capacitors $C1$, $C2$, and $C4$. Resistors $R1$ and $R2$ are ac bypassed and do not appear in the ac circuit. Tunnel diode $CR1$ is replaced by negative resistor R_D and capacitor C_D. At the particular operating frequency (about 1 mc), the lead inductance and resistance of the diode are negligible and are not shown. This circuit is the equivalent circuit of the amplifier and can be used to determine its behavior at any frequency.

c. At the operating frequency the parallel tuned circuit formed by capacitors C_D and $C3$ and coil $L1$ displays infinite impedance. The equivalent circuit is reduced to that shown in Fig. 4–10C. The analysis for this circuit is covered in paragraph 3–25. For stability the resistance of R_L must be less than the negative resistance R_D.

4–14. RF (100 mc) Amplifier

a. Figure 4–11A shows a 100-mc amplifier operating between two 50-ohm transmission lines. Tunnel diode $CR1$ has the following characteristics:

1. Peak current (I_p) = 1.0 ma
2. Valley current (I_v) = 0.15 ma
3. Lead inductance (L_s) = 5×10^{-9} h
4. Internal resistance (R_s) = 2 ohms
5. Diode capacitance C_D = 5 mmf
6. Negative resistance $(-R_D)$ = -143 ohms
7. Self-resonant frequency (f_s) = 985 mc
8. Resistance cutoff frequency (f_r) = 1.9 kmc

At quiescence the diode is biased at the inflection point which is at 125 mv and 0.75 ma.

b. The ac circuit of the amplifier is shown in Fig. 4–11B. Resistor R_s is the signal generator output resistance. Resistor R_Z is the series impedance-matching resistance. Coil $L1$, capacitor C_D, plus lead inductance, and stray capacitance provide selectivity by forming a series circuit resonant at 100 mc.

c. It can be shown by using the formula in paragraph 3–18 that, at the 100-mc operating frequency, the negative resistance is reduced from -143 ohms (*a*6 above) to an effective value of -118 ohms. Note that the resistances at the ends of the transmission lines total 100 ohms; add the lead

resistance (a4 above) of 2 ohms and the total positive resistance is 102 ohms. To obtain maximum gain, a resistance of 16 ohms must be added in series so that the total resistance is 118 ohms to match the effective negative resistance of 118 ohms. The actual dc resistance is slightly higher than 118 ohms because of the dc resistance of the transmission lines; therefore oscillation does not occur (par. 3–16).

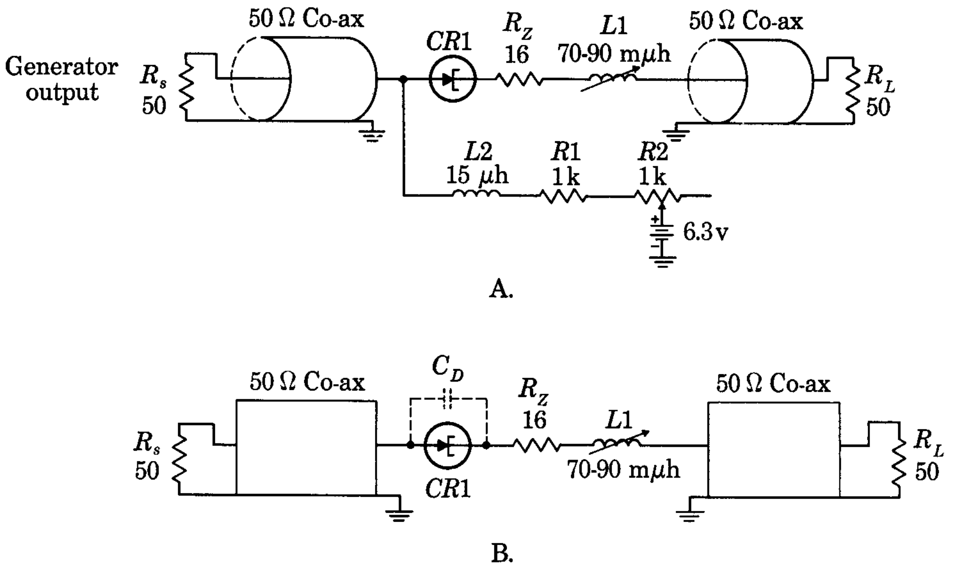


FIG. 4–11. RF (100 mc) amplifier and its ac circuit (Adapted circuit, courtesy General Electric Company)

d. The dc circuit consists of RF decoupling choke $L2$ (Fig. 4–11A) which has a reactance of 10,000 ohms at 100 mc, and dc dropping resistors $R1$ and $R2$. Variable resistor $R2$ permits adjustment of the bias voltage across the diode to 125 mv and bias current to 0.75 ma (a above).

e. The circuit provides a gain of 32 db at 100 mc with a bandwidth of 20 mc. If inductance $L1$ is increased to give a higher Q (selectivity) to the series resonant circuit, the gain can be increased to 40 db but the bandwidth is nonsymmetrical and is reduced to 8 mc. A practical application for this circuit would be as a low-level repeater in a transmission line. Note that the circuit is bilateral; that it can amplify a signal in either direction.

4–15. RF (30 mc) Amplifier

a. Figure 4–12 shows a 30-mc amplifier operating between two 50-ohm transmission lines. The characteristics of the tunnel diode are given

in paragraph 4–14*a*. Tunnel diode *CR*1 is biased at the inflection point. The amplifier produces a gain of 32 db at the operating frequency and provides a bandwidth of 10 mc.

b. RF choke *L*2, having a reactance of 10,000 ohms at 30 mc, isolates the ac components of the circuit from the power supply. Resistor R_s is the signal generator output resistance. Capacitor *C*1, in parallel with the diode internal capacitance, forms a series circuit with coil *L*1 resonant at 30 mc. Resistor *R*1 suppresses parasitic oscillations. Resistor R_L is the load resistor. Resistors *R*2 and *R*3 reduce the battery voltage to the required value for biasing the diode at the inflection point.

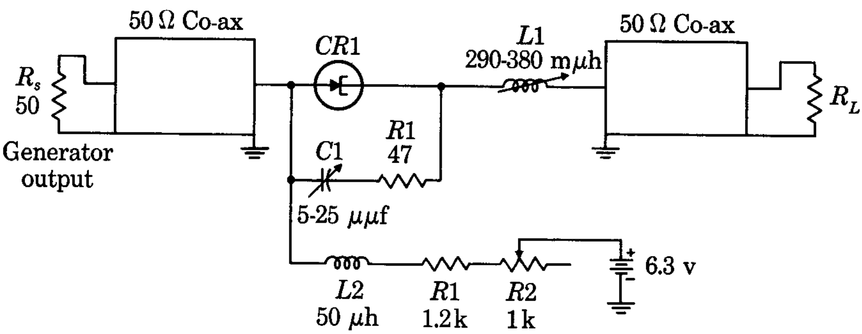


FIG. 4–12. RF (30 mc) amplifier (Adapted circuit, courtesy General Electric Company)

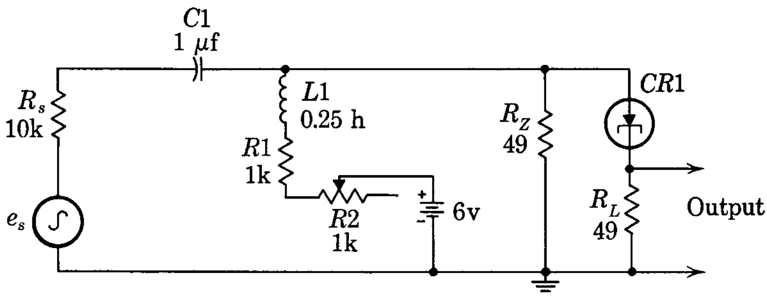
c. Because the internal diode capacitance alone is insufficient to resonate with coil *L*1 at the operating frequency and also provide the required bandwidth, capacitor *C*1 must be added to the circuit. If capacitor *C*1 alone were added directly across the tunnel diode, the resultant loop formed by the two components would oscillate parasitically at approximately 350 mc. Resistor *R*1, added in series with capacitor *C*1, suppresses the parasitic oscillation by reducing the resistance cutoff frequency of the diode from 1.9 kmc (par. 4–14*a*8) to less than 350 mc.

d. Capacitor *C*1 and resistor *R*1 also reduce the tunnel diode negative resistance of −143 ohms (par. 4–14*a*6) to an effective value of −100 ohms. With resistors R_s and R_L plus line resistance adding to slightly more than 100 ohms, the series circuit will be stable (nonoscillatory) (par. 3–16).

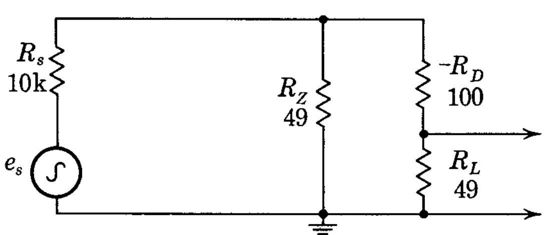
4–16. Tunnel Diode Audio Amplifier

a. Figure 4–13A shows a tunnel diode audio amplifier. The signal source is represented by signal generator e_s having an internal resistance R_s; this arrangement could represent the output voltage and impedance of a tran-

sistor amplifier. Capacitor $C1$ blocks dc voltage from the signal source. Coil $L1$ isolates the dc bias circuit from the ac circuit. Resistor R_Z prevents oscillation. The presence of diode $CR1$ causes an amplified signal to appear across load resistor R_L. Resistor $R1$ and variable resistor $R2$ drop the dc voltage to the required value and permit biasing of diode $CR1$ at the inflection point.



A. Audio amplifier



B. Equivalent circuit

FIG. 4–13. Tunnel diode audio amplifier and its equivalent circuit

b. The ac equivalent circuit is shown in Fig. 4–13B. Tunnel diode $CR1$ is represented by its negative resistance of -100 ohms. At audio frequencies the internal capacitance of the diode is negligible and is not shown. The total resistance in the diode branch is -51 ohms ($49 - 100$). Resistor R_Z, in parallel with the diode branch, prevents oscillation because its value (49 ohms) is less than the absolute value (51 ohms) in the diode branch. In a parallel arrangement such as this, the tunnel diode is short-circuit stable (par. 3–27). Note that this circuit is similar to the circuit shown in Fig. 3–28 and discussed in paragraph 3–32. However, the functions of impedance-matching resistor R_Z and load resistor R_L have been interchanged.

4–17. Summary

a. The tunnel diode exhibits shot noise and thermal noise. The shot noise is generally the result of the tunneling of electrons; the thermal noise is due to bulk resistance.

b. The tunnel diode exhibits low noise when operating at frequencies above 25 kc and below $0.707f_r$ (resistance cutoff frequency).

c. A plot of the negative resistance of the tunnel diode versus the bias voltage indicates the nonlinearity of the negative resistance. This curve can be used to indicate the distortion introduced during amplification.

d. The inflection point of the diode characteristic is the point of minimum value of negative resistance.

e. The peak-current of a tunnel diode may rise or fall with increasing temperatures, depending upon the type of impurities used in the semiconductor material; the valley current always increases with temperature.

f. Tunnel diode amplifiers must be biased from well-regulated voltage sources to avoid variations in negative resistance and therefore in amplifier gain.

g. Most tunnel diode amplifiers are biased at the inflection point.

h. Tunnel diode RF amplifiers (par. 4–14 and 4–15) are easily adapted for use in transmission-line repeaters because they are bilateral (amplify in either direction).

Chapter 5

CASCADING TUNNEL DIODE AMPLIFIERS

SECTION I. GENERAL

5–1. Two-Port and One-Port Devices

a. The transistor and the electron tube are referred to as two-port devices. Each *set* of terminals for input or output use is referred to as a *port,* even if one of the terminals is common to the input and the output. The word *port* is derived from Latin and is used in English to mean an entrance or an exit. The tunnel diode, then, is a one-port device, the one port being used for input *and* output.

b. In the case of the two-port devices mentioned, isolation of the input circuit from the output circuit is, for most applications, inherent in the structure of the device. Only in the case of very high-frequency applications is the feedback of energy from the output circuit to the input circuit through interelectrode capacitance sufficient to cause oscillation. Even in these cases minor circuit adjustments are necessary to prevent oscillation in a given two-port device amplifier or a series of cascaded amplifiers.

c. In the case of a *single* tunnel diode amplifier stage, the technique for preventing oscillation is a relatively simple one. If used in a series amplifier, it is necessary only to ensure that the total positive resistance of the circuit is larger than the effective negative resistance of the circuit. If used in a parallel amplifier, it is necessary only to ensure that the total positive resistance of the circuit is smaller than the effective negative resistance of the device. When more than one tunnel diode amplifier stage is necessary to achieve the desired overall gain, it is not a simple matter of coupling the output of one stage directly into the other stage. Even if each stage is designed to be stable in accordance with the conditions set forth above for a series or a parallel amplifier, unless special precautions are taken (par. 5–2), energy from a high-level stage can flow back to a low-level stage and cause oscillation. This situation occurs regardless of the operating frequency of the amplifier stages.

5–2. Isolation

In order to cascade tunnel diode amplifier stages, the circuit arrangement or special components used must ensure the flow of energy from low-level to high-level power stages while *isolating* the output of higher-level stages from lower-level stages. In other words, the flow of energy through the amplifiers must be unidirectional. Isolating methods or devices used to produce unidirectional cascaded tunnel diode amplifiers are as follows:

a. Matched transmission line technique (par. 5–3 and 5–4)
b. Quarter-wave line technique (par. 5–5 through 5–8)
c. Hall-effect and skew isolators (par. 5–9 through 5–15)
d. Hybrid junctions (par. 5–16 through 5–19)
e. Ferrite circulators (par. 5–20 through 5–25)

SECTION II. MATCHED TRANSMISSION LINE TECHNIQUE

5–3. General

a. Figure 5–1A shows a transmission line into which power is introduced at one end of the line by a signal source (e_s). The power is transmitted over the line and absorbed by the load resistor (R_L) at the other end of the

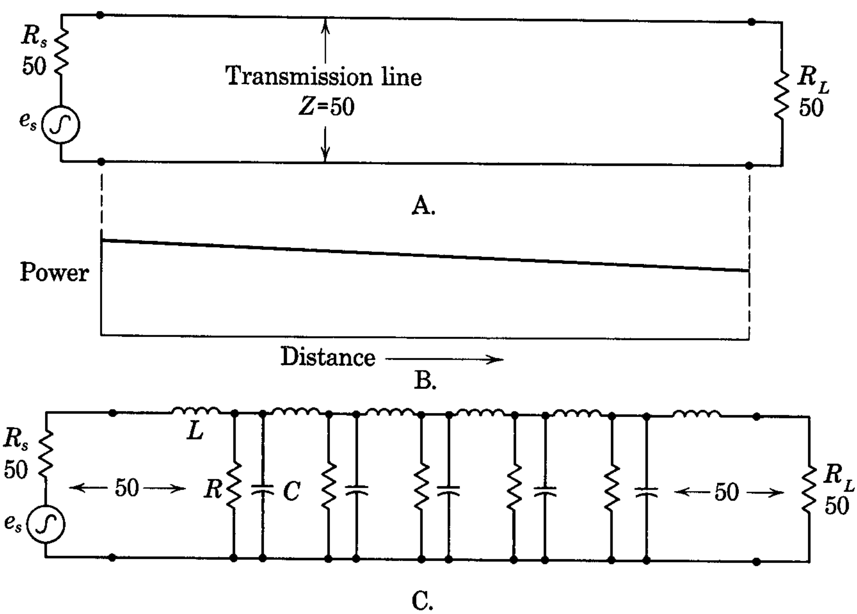


FIG. 5–1. Transmission line, power distribution, and equivalent circuit

line. If the internal resistance (R_s) of the signal source, the characteristic impedance (Z) of the line, and the resistance of the load are equal in value, none of the power introduced will be *reflected* from the load end back to the source. If power were reflected, standing waves would result by the interaction of transmitted and reflected power. Figure 5–1B shows the reduction in power as it is transmitted over the line. Note, however, that because of the *matched* condition wherein $R_s = Z = R_L$, no power is reflected.

b. In Fig. 5–1C the transmission line has been replaced by its electrical equivalent of lumped parameters. To the electrical signal the line looks like a series of inductances and parallel capacitors and resistors. Reduced power is delivered to the load because some of the power from the signal source is absorbed by the resistors. If the positive resistance were replaced by negative resistance (par. 5–4), power and voltage gain would result.

5–4. Cascading Matched Amplifiers

a. To apply the matched transmission line technique to cause a unidirectional signal flow in a series of cascaded tunnel diode amplifiers, it is necessary to design the circuit so that the signal *sees* the *same* impedance in moving from the source to the load. A simplified version of a method proposed to achieve the desired results is discussed below.

b. This discussion considers the load end first and works toward the signal source. Throughout the discussion the tunnel diode is represented as an *ideal* negative resistance; i.e., its inherent junction capacitance, lead inductance, and resistance are ignored.

1. Figure 5–2A shows a load resistor (R_L) of 50 ohms in series with a negative resistance (R_{D1}) of 45 ohms. The input resistance to this combination is 5 ohms:

$$R_{\text{in}} = 50 - 45 = 5 \text{ ohms}$$

2. By placing -5.55 ohms in parallel (Fig. 5–2B) with the series combination (1 above) the input resistance will be $+50$ ohms, calculated as follows:

$$R_{\text{in}} = \frac{-5.55(50 - 45)}{-5.55 + 50 - 45}$$

$$R_{\text{in}} = \frac{-5.55(5)}{-0.55} = +50 \text{ ohms}$$

3. The input resistance ($+50$ ohms) of this arrangement (2 above) can be used as the load for a similar combination of negative resistances (Fig. 5–2C). The input resistance again will be $+50$ ohms. The same pattern can be used for a number of stages, in which the input of one stage acts as

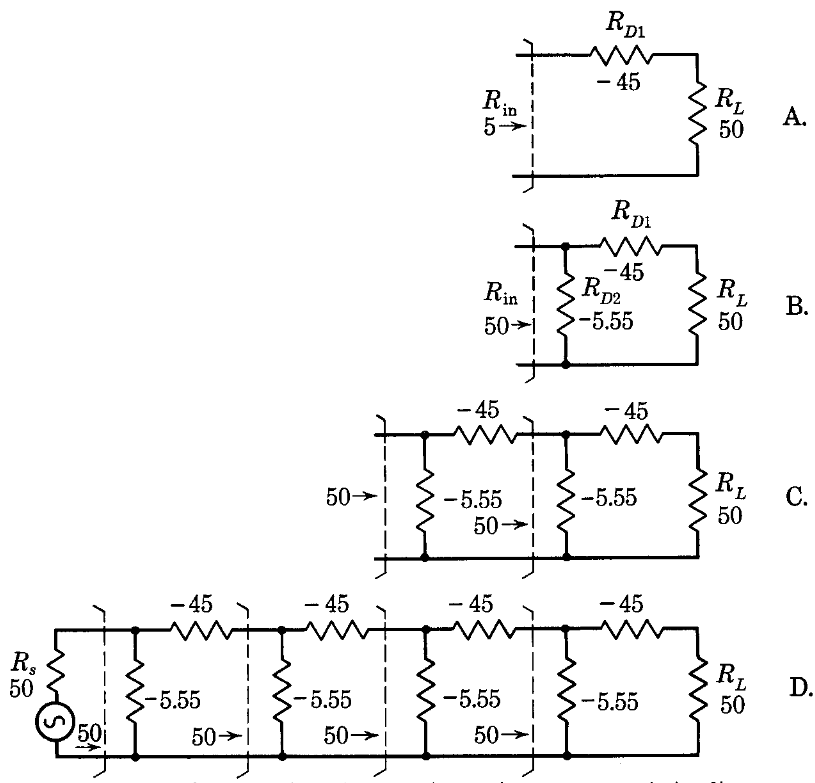


FIG. 5–2. Construction of a *negative-resistance transmission line*

the load of the preceding stage. Four such stages are shown in Fig. 5–2D. Note that the source impedance is 50 ohms and *sees* 50 ohms. Each stage *sees* 50 ohms, and the load is 50 ohms. This condition is the equivalent of the transmission line arrangement discussed in paragraph 5–3.

c. A simplified circuit of cascaded tunnel diode amplifiers using this technique is shown in Fig. 5–3A. Diodes $CR1$ through $CR4$ provide the -45 ohms in their respective stages; diodes $CR5$ through $CR8$ provide the -5.5 ohms in their respective stages. Capacitors $C1$ through $C4$ are dc blocking capacitors. Each tuned circuit consisting of a capacitor C_p and an inductance L_p provides frequency selectivity. At resonance the tuned circuit impedance is infinite; the equivalent circuit (*b* above) will not be affected by the presence of the tuned circuits. It can be shown that the voltage gain of each stage equals 10. Figure 5–3B (not drawn to scale) indicates the voltage level at each stage. Each stage multiplies the previous gain by 10, so that the overall voltage gain is 10,000 times.

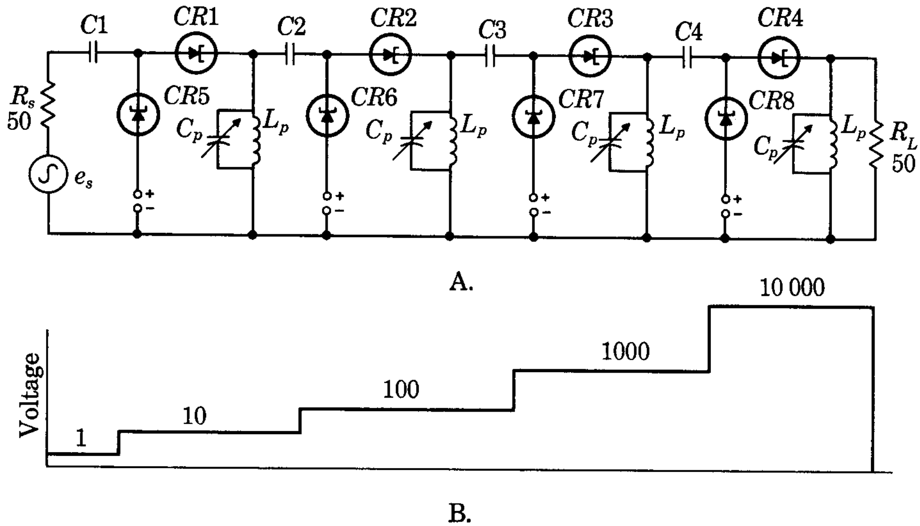


FIG. 5–3. Cascaded tunnel diode amplifiers using the matched impedance line technique

d. Practical circuits employing this technique have not been sufficiently investigated. However, advanced theoretical analysis indicates possible application at very high frequencies only. The main difficulty is that tunnel diodes are not *ideal* negative-resistance devices. Isolation techniques using successfully tested methods are covered in subsequent sections of this chapter.

SECTION III. MATCHED AMPLIFIERS USING QUARTER-WAVE TRANSMISSION LINE

5–5. General

A transmission line, cut to one quarter of the wavelength of the frequency being transmitted, can be properly terminated at each end to match two tunnel diode amplifiers. The matched condition results in a unidirectional flow of energy between the cascaded amplifiers. A discussion of the properties of matched quarter-wave lines is given in paragraphs 5–6 and 5–7. A practical circuit of cascaded tunnel diode amplifiers is discussed in paragraph 5–8.

5–6. Quarter-Wave Line

a. Figure 5–4A shows two parallel wires cut to a quarter of the wavelength of the frequency arbitrarily chosen for discussion purposes. The frequency is 30 mc; one quarter of its wavelength is 2.5 meters (over 7.5 ft). The characteristic impedance of the line is designated Z.

b. Figure 5–4B shows the same quarter-wave line terminated in a load resistance (R_L) equal to its characteristic impedance (Z). The quarter-wave line is fed from a 30-mc signal source whose internal resistance (R_s) equals Z. With proper matching at each end of the line, the voltage (E) and current (I) distribution along the line is as indicated. Note that there are no standing waves of voltage or current that would be caused by reflections from an unmatched load. All the energy transmitted by the source is absorbed by the load.

c. Figure 5–4C shows the same quarter-wave line terminated by two negative resistances (R_{D1} and R_{D2}). The value of each is -50 ohms. The line is fed from a 30-mc source having an internal resistance (R_s) of 50 ohms

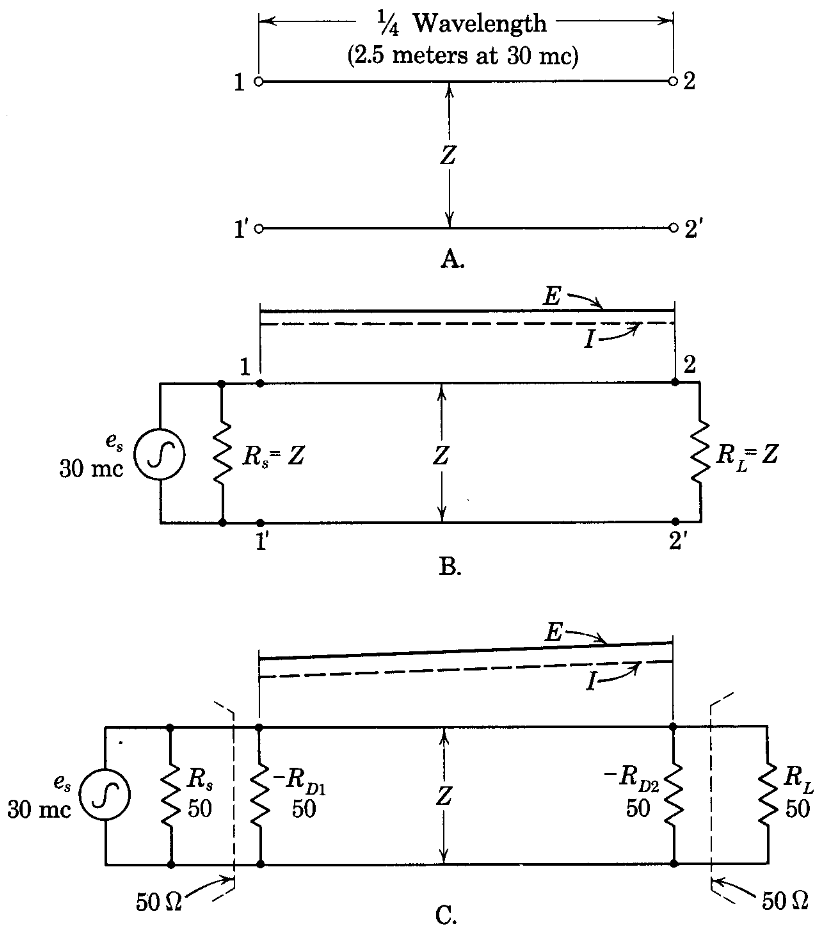


FIG. 5–4. Quarter-wave transmission line and two possible matched conditions

and is terminated in a 50-ohm load (R_L). The voltage (E) and current (I) distribution along the line is also shown. Note that the *rise* in current and voltage is due to the presence of the *negative resistances*. Note also that there are no standing waves (or portions of a standing wave) that would be caused by reflections. A matched condition must exist as in the case discussed in *b* above. For this condition to exist, the 50-ohm resistance must be *looking* into a 50-ohm resistance; the 50-ohm load must be *looking* back into a 50-ohm resistance; i.e., the ratio of voltage (E) to current (I) at the input and output must be 50. By mathematical analysis beyond the scope of this book, it can be shown that this condition can be accomplished by choosing the proper values of negative resistance ($-R_D$) and characteristic impedance (Z) for the particular values of load and source resistances. A qualitative (nonmathematical) analysis of the phenomena involved is given in paragraph 5–7.

5–7. Open-End Quarter-Wave Line

The matched quarter-wave line using negative-resistance terminations (par. 5–6) can be understood by considering the properties of an open-end quarter-wave line.

a. Figure 5–5A shows a quarter-wave line being fed by a signal source of the frequency for which the line is cut. Because of the open end, standing waves of voltage (E_1) and current (I_1) as indicated are established. Note that the open end *cannot* absorb energy from the source because of the *zero* current at the open end. The source *sees* a low resistance at terminals 1–1′ because of the low ratio of zero voltage to maximum current at 1–1′. The line acts like a *series* resonant circuit. Figure 5–5B shows the same condition with the source now located at terminals 2–2′, and terminals 1–1′ are open. The voltage (E_2) and current (I_2) distribution are as indicated.

b. If it were possible to transmit from both ends simultaneously while maintaining *both* ends open, the current and voltage distribution over the quarter-wave line would be as indicated in Fig. 5–5C. Here, the voltage line ($E_1 + E_2$) is the result of adding the voltage lines shown in Figs. 5–5A and B. The same procedure is used to obtain the current line $I_1 + I_2$. Note that the input resistances at either end of the line are no longer equal to zero, but have a finite positive value equal to the ratio of voltage to current.

c. Figure 5–6A shows how the condition cited in *b* above can be brought about by using the negative resistance of properly terminated tunnel diode amplifiers. At terminals 1–1′, a tunnel diode having a negative resistance ($-R_D$) of 50 ohms paralleled by a positive resistance (R_s) of 50 ohms is

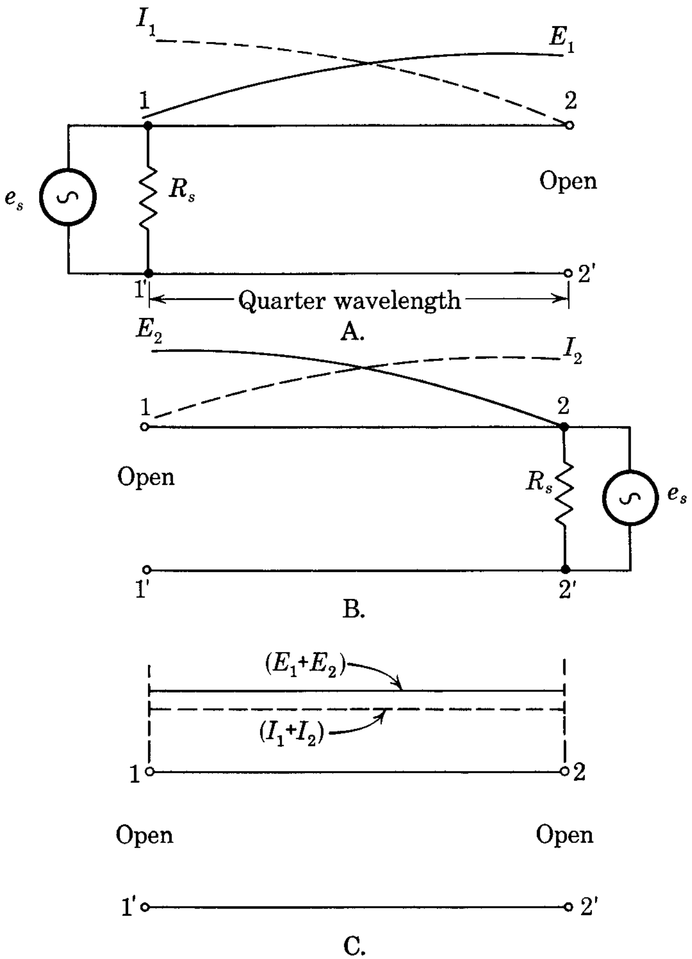


FIG. 5–5. Quarter-wave lines, single-end open, and double-end open

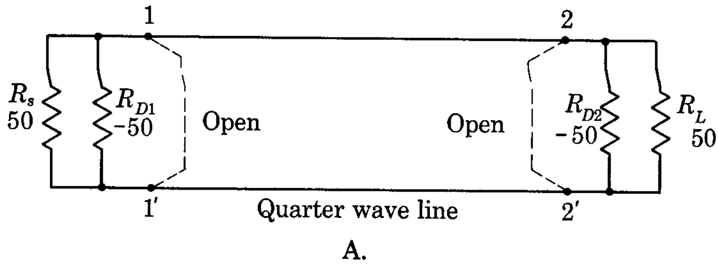
shown. The resistance (R) of the network at terminals 1–1′ is infinite (open circuit), calculated as follows:

$$R = \frac{50(-50)}{50 + (-50)}$$

$$= \frac{-2500}{0}$$

$$R = \text{Infinite } (\infty)$$

The same prevails at terminals 2–2′.



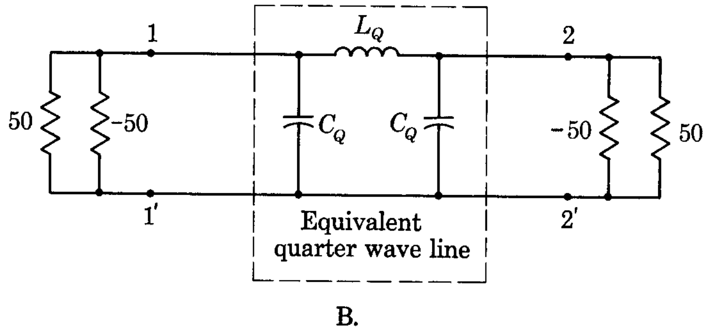


FIG. 5–6. Positive resistances paralleled with negative resistances to achieve double-end open condition

d. By using the proper value of characteristic impedance (Z), the ratio of voltage to current at the terminals (Fig. 5–5C) can be made equal to 50 ohms. Thus resistance R_s (Fig. 5–6B) representing a source resistance of 50 ohms would *look* into 50-ohm terminals, and load resistance R_L (50 ohms) would *look* back at 50 ohms. A matched condition would prevail.

e. The quarter-wave line at 30 mc would be more than 7.5 ft long and would be awkward. In its place an equivalent quarter-wave line of lumped constants could be used; the equivalent is represented by the network formed by coil L_Q and capacitors C_Q. A practical circuit employing the quarter-wave line is discussed in paragraph 5–8.

5–8. Matched Amplifiers

a. Two cascaded tunnel diode amplifiers matched by a quarter-wave line for unidirectional energy flow are shown in Fig. 5–7. Coil L_Q and capacitors C_Q represent a 30-mc quarter-wave line. A signal (e_s) is fed to the amplifier by the signal source whose resistance (R_s) is 50 ohms. Transformer $T1$ couples the signal to tunnel diode $CR1$; transformer $T1$ is used for impedance matching. The negative resistance (of $CR1$) reflected to the line by $T1$ is slightly *larger* than the positive resistance of R_s to ensure non-

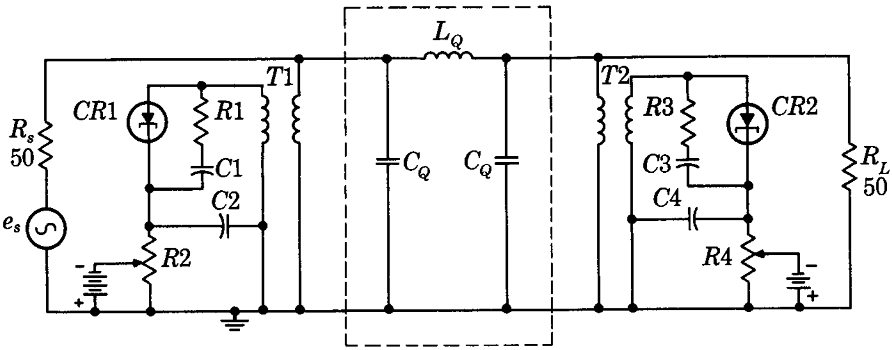


FIG. 5-7. Tunnel diode amplifiers cascaded through an equivalent quarter-wave line

oscillation. Resistor $R2$ adjusts the bias for diode $CR1$. Capacitor $C2$ decouples the bias source from the amplifier. At higher frequencies the leakage reactance of transformer $T1$ prevents coupling of diode $CR1$ to its load (R_s and the line). To prevent oscillation at these frequencies, capacitor $C1$ and resistor $R1$ act as a load; the reactance of $C1$ is very low at the higher frequencies. At lower frequencies the low resistance of the transformer shunts the diode to prevent oscillation. The presence of diode $CR1$ causes amplification of the signal.

b. On the other side of the quarter-wave line is a second amplifier which further amplifies the signal. The correspondingly positioned circuit elements perform the same function in the second amplifier, except that resistor R_L is the load resistor.

c. Assuming ideal transformers, the ac equivalent circuit of this amplifier at resonance is similar to the circuit shown in Fig. 5-6B. The only differ-

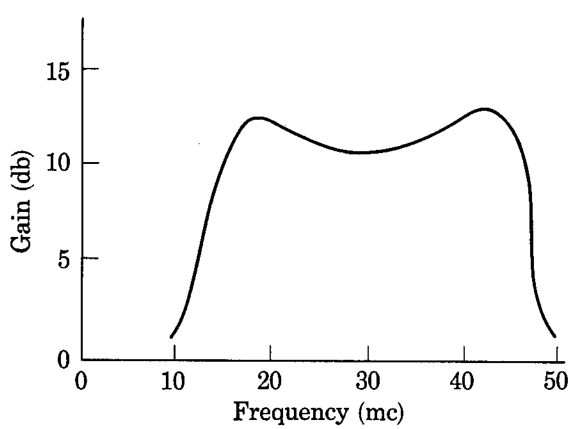


FIG. 5-8. Chart of gain versus frequency

ence is that the ac negative resistance is slightly higher than that shown.

d. A chart of gain (db) versus frequency (mc) of a practical amplifier is shown in Fig. 5–8. A gain of approximately 9 db occurs at the 30-mc operating frequency with a relatively flat bandwidth of 10 mc, centered at 30 mc. Peaks occur at 17 and 40 mc. At these frequencies the *line* is sufficiently removed from a quarter-wave length to cause reflections of energy and increased gain.

SECTION IV. HALL-EFFECT ISOLATORS

5–9. General

a. In essence, the Hall effect (par. 5–10) refers to the phenomenon of the occurrence across a conductor of a voltage that is at right angles to the externally applied voltage when the conductor is subjected to an external magnetic field. The Hall effect was first observed in 1880 and was not considered for practical application until the intensive interest in semiconductor materials developed late in the 1940's. With respect to the particular material, the Hall voltage is directly proportional to the ratio of the mobility of the current carriers to the conductivity of the material. In the case of conductors such as copper, silver, and aluminum, the conductivity is so high that this ratio is low, with the result that the Hall voltage is almost insignificant. In semiconductors the conductivity is so much lower (higher resistance) that the ratio of mobility to conductivity is relatively large, and the Hall voltage becomes appreciable in magnitude. One of the first practical applications of the Hall effect was in the measurement of the mobility of current carriers in *p*- and *n*-type semiconductors. This application is covered extensively in the literature (Appendix A).

b. The Hall voltage measured is also proportional to the magnitude of the current flow through the material and the intensity of the externally applied magnetic field. Because the current flow in insulators such as rubber and polystyrene is very small, the Hall voltage in the case of insulators is difficult to detect.

c. This section covers the following topics:

1. Hall-effect phenomenon (par. 5–10)
2. Hall-effect gyrator (par. 5–11)
3. Hall-effect isolator (par. 5–12)
4. Skew isolator (par. 5–13)
5. Unidirectional tunnel diode amplifiers using Hall-effect isolators (par. 5–14 and 5–15)

5-10. Hall Effect

The Hall effect is most pronounced in semiconductor materials. In addition to the type of material, the shape of the material is also important. A relatively thin wafer of material is most desirable to make the effect significant.

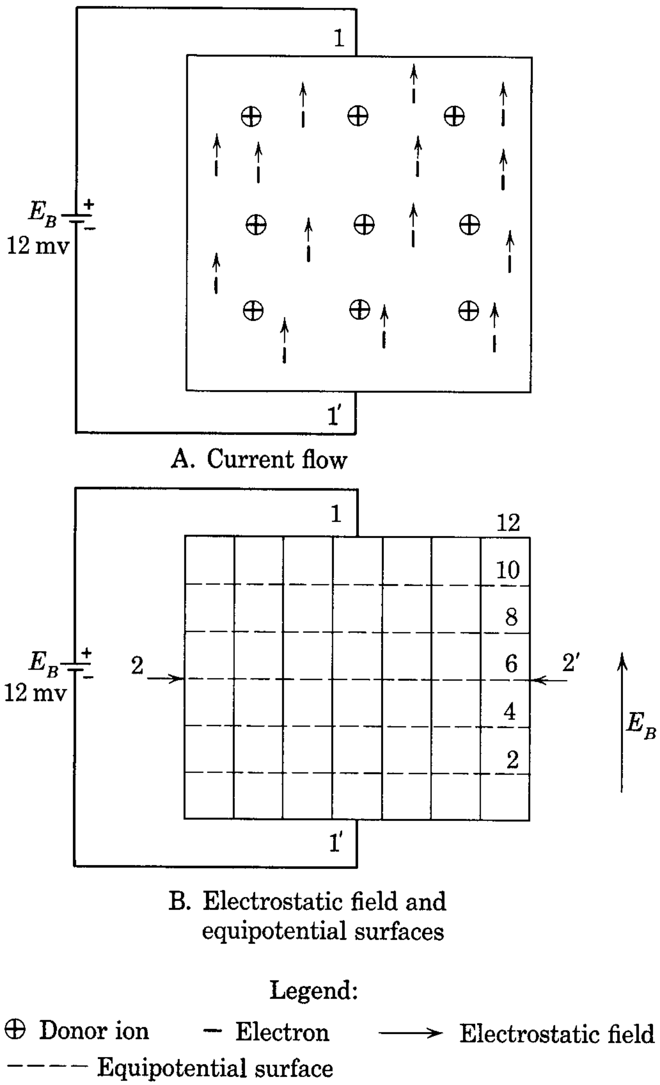


FIG. 5-9. Current flow and equipotential surfaces in n-type semiconductor

a. Applied Voltage Only. 1. Figure 5–9A shows a thin square wafer of n-type semiconductor across which a battery voltage of 12 mv is applied at terminals 1–1′. The wafer is *not* subjected to an external magnetic field. The current carriers (negative electrons) flow from the negative potential to the positive potential in approximately straight lines that are uniformly spaced. Figure 5–9B shows the same wafer under the same conditions except that the electrical field within the material and the equipotential surfaces are shown. The electrical field (solid-line arrows) is in the direction that an electron would move if placed within the electrical field. The equipotential surfaces (dashed lines) are at right angles to the direction of the electrical field. The potential on each equipotential surface is also indicated. The electrical field is shown externally as vector E_B.

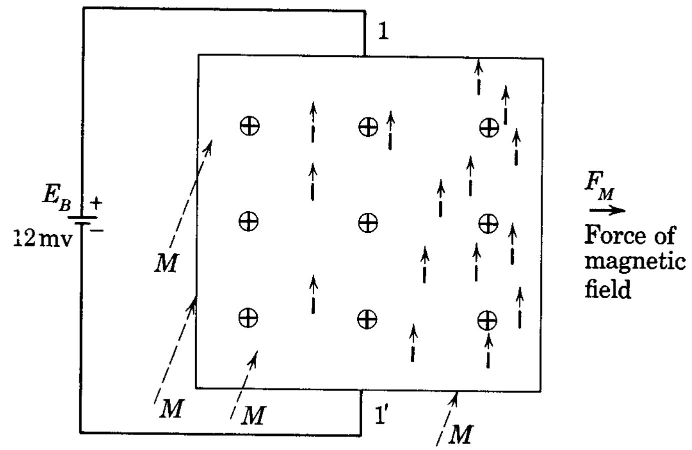
2. In Fig. 5–9B two terminals 2–2′ are attached at exactly opposite points as indicated. If a voltmeter were attached to terminals 2–2′, there would be no difference of potential because the two terminals are on the same 6-mv equipotential surface.

b. Applied Voltage and Magnetic Field. 1. Figure 5–10A shows the same wafer with the same battery voltage applied at terminals 1 and 1′. Simultaneously a magnetic field (dashed-line arrows terminated in the letter M) is applied at right angles to the direction of the electrical field and the current flow. The magnetic field exerts a force F_M on the electrons in the direction shown. Note that the electrons are displaced to the right. The displacement to the right creates a transverse electrical field between the negative electrons and the fixed positive donor ions on the left. The displacement continues until the force of the transverse electrical field is equal and opposite to the magnetic force F_M.

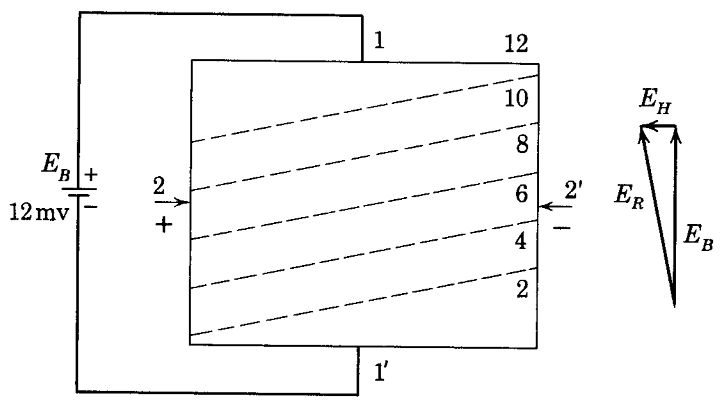
2. The voltage (E_H) created by the transverse electrical field is shown as a vector in Fig. 5–10B. The voltage (E_B) of the battery is also shown by a vector. The resultant electrical field (E_R) is the sum of E_B and E_H. The equipotential lines shown within the wafer are at right angles to the resultant electrical field E_R.

3. If a voltmeter is now placed across terminals 2 and 2′, a voltage will be indicated. The voltage will be equal to E_H, which is called the Hall voltage, and will have the polarity indicated. Note that the Hall effect *rotates* the equipotential surfaces and creates a transverse voltage. Compare B of Figs. 5–9 and 5–10.

4. For the above discussion, a wafer of n-type semiconductor was used. Use of p-type semiconductor would result in the rotation of the equipotential surfaces in the opposite direction, so that terminal 2 would be negative and terminal 2′ would be positive.



A. External magnetic field applied



B. Rotation of equipotential surfaces

Legend:

M ---→ Magnetic field

FIG. 5–10. Rotation of equipotential surfaces caused by application of external magnetic field

5–11. Hall-Effect Gyrator

The Hall-effect gyrator (Fig. 5–11) is a device that employs the Hall effect (par. 5–10) to bring about a *nonreciprocal* current flow relationship between two circuits connected to the same semiconductor wafer. In this case a wafer of n-type semiconductor, immersed in a magnetic field at right

angles to the direction of current flow, is used. The circuit connected to terminals 1–1′ is referred to as the *primary* circuit; that connected to terminals 2–2′ is referred to as the *secondary* circuit.

a. In Fig. 5–11A a voltage source is placed in the primary circuit as indicated. The Hall effect rotates the equipotential surfaces (dashed lines) as indicated and creates a voltage at terminals 2–2′. This voltage causes a current to flow through resistor R_s. Note that the electron current flow (solid-line arrows) in the primary and in the secondary circuit is in the *counterclockwise* direction.

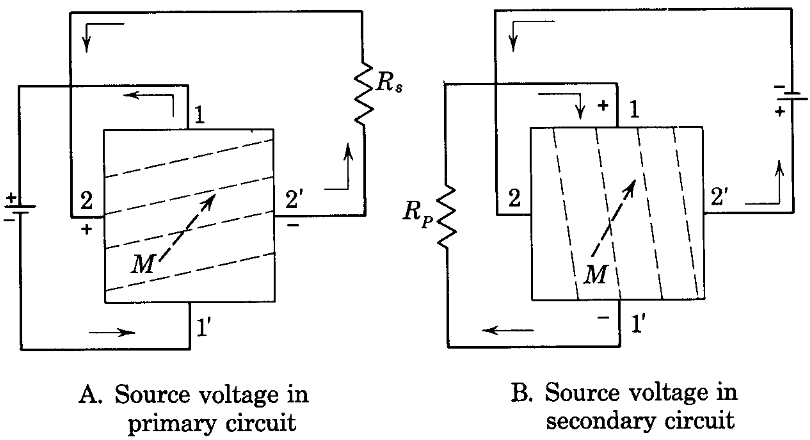


FIG. 5–11. Hall-effect gyrator

b. In Fig. 5–11B the voltage source is placed in the *secondary* circuit with the polarity arranged to cause electron current to flow in the counterclockwise direction in the secondary circuit. The Hall effect again rotates the equipotential surfaces, as indicated, to cause a voltage to appear at terminals 1–1′ with the indicated polarity. Note that electron current flow in the primary circuit is in the *clockwise* direction.

c. Depending on the direction that electrical energy is transmitted, left-to-right or right-to-left, the currents in the circuits may be in the same direction (*a* above), or in opposite directions (*b* above), respectively. For this reason, the device is referred to as *nonreciprocal*. Because of the 180° reversal of current flow between the two conditions, the device rotates or *gyrates* directions of current flow and is called a Hall-effect *gyrator*.

d. By using feedback (shunt) resistors, the gyrator can be transformed into an *isolator* that causes energy to flow in one direction only (par. 5–12).

5–12. Hall-Effect Isolator

a. The Hall-effect gyrator (par. 5–11) is shown in Fig. 5–12 as a square wafer with terminals 1–1′ and 2–2′ attached as indicated. The magnetic field to which it is exposed is represented in the center of the square as a dashed-line arrow terminated in the letter M. Except for the addition of feedback resistors R_{F1} and R_{F2}, the primary and secondary circuits are identical with those shown in Fig. 5–11B.

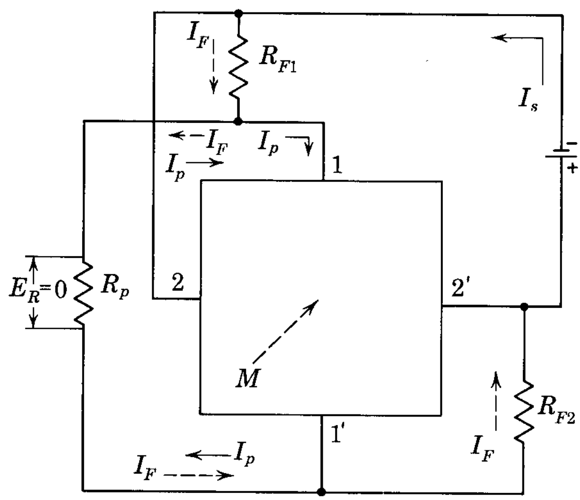


FIG. 5–12. Hall-effect isolator

b. The battery in the secondary circuit causes current I_s to flow in the direction shown. The Hall effect causes a current I_p to flow in the primary circuit in the direction shown. A portion of the secondary current (I_s) is shunted to the primary circuit through feedback resistors R_{F1} and R_{F2}. The feedback current is designated I_F. If the values of resistors R_{F1} and R_{F2} are properly selected, I_F can be made to equal I_p. Since these currents flow in opposite directions in resistor R_p, the net current flow is zero, the voltage drop E_R is zero, and no battery energy is transferred from the secondary circuit to the primary circuit. If the battery were reversed in the secondary circuit, all current directions would be reversed and there still would be no transfer of energy from the secondary to the primary circuit. If the positions of the battery and the load resistor (R_p) were interchanged, there would be current flow through the load resistor even with the presence of the feedback resistors (R_{F1} and R_{F2}). In this condition the current

through the resistor (in the secondary circuit) transferred through the gyrator would be in the same direction as the current through the resistor coupled through the feedback resistors from primary to secondary circuit.

c. The combined effect of the gyrator and the feedback resistors is to cause the transfer of electrical energy in *one* direction only, from primary to secondary circuit. The combination is referred to as a Hall-effect isolator. Practical germanium Hall-effect isolators have a forward loss of approximately 6 db minimum, and a reverse loss of 75 db, with impedances ranging from 10 to 1000 ohms. Hall-effect isolators made from indium antimonide have a forward loss as low as 7.5 db, and a reverse loss as high as 100 db. However, the impedance is usually as low as 1 ohm. The theoretical frequency limit of Hall-effect isolators made of semiconductor materials is 10^{12} cycles per second. In practice this frequency is substantially reduced by lead inductance and stray capacitance.

d. Use of the Hall-effect isolator in cascading tunnel diode amplifiers is covered in paragraph 5–14.

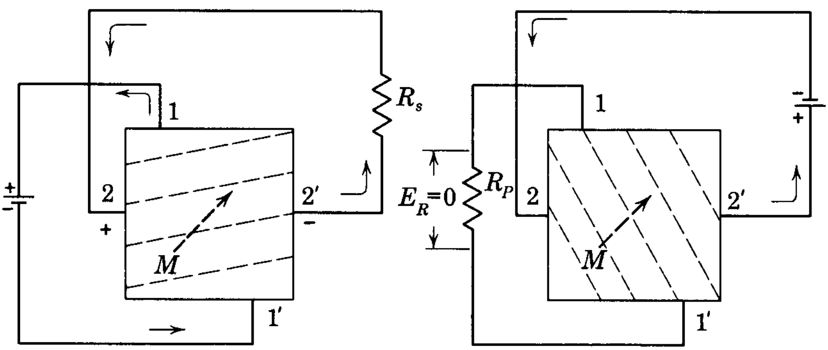
5–13. Skew Isolator

The skew isolator is made of a thin wafer of *n*- or *p*-type semiconductor material. Like the Hall-effect gyrator (par. 5–12) the skew isolator depends for its isolating (unidirectional) properties on the Hall effect. Figure 5–13 shows a skew isolator made of *n*-type semiconductor material. Note that terminals 2–2′ are directly opposite each other and in the center of their respective sides. Terminals 1–1′ are offset in opposite directions from the center line. Because of the *nonsymmetrical* placement of the terminals, the device is called a *skew* isolator.

a. Figure 5–13A shows a battery connected across terminals 1–1′ so that counterclockwise current flows in the primary circuit. The presence of the external magnetic field M rotates the equipotential surfaces (Hall effect) so that a voltage appears at terminals 2–2′ and counterclockwise current flows in the secondary circuit. If the polarity of the battery in the primary circuit were reversed, both current directions would be reversed. Thus energy can be transferred by the Hall effect from the primary circuit to the secondary circuit.

b. Figure 5–13B shows a battery connected across terminals 2–2′ so that counterclockwise current flows in the secondary circuit. The presence of the external magnetic field rotates the equipotential surfaces as indicated. Note that terminals 1–1′ are located on the same internal equipotential surface; there is, therefore, no voltage difference across terminals 1–1′ and no current flows through the primary circuit. The voltage (E_R) across primary resistor R_p is zero and no energy is transferred to the pri-

mary circuit. If the battery in the secondary circuit were reversed in direction, the equipotential surfaces would rotate in the opposite direction; terminals 1–1′ would no longer be on the same equipotential surface, and current would flow in the primary circuit. To use the device as an isolator, therefore, the current in the secondary circuit must be maintained in the counterclockwise direction. The current, however, may vary in magnitude, and it may *carry* an ac signal without transferring ac or dc energy to the primary circuit.



A. Source voltage in primary circuit

B. Source voltage in secondary circuit

FIG. 5–13. Skew isolator

c. Use of the skew isolator in cascading tunnel diode amplifiers is discussed in paragraph 5–15.

5–14. Cascaded Amplifiers Using Hall-Effect Isolators

a. Transformer Coupling. Figure 5–14 shows two cascaded tunnel diode amplifier stages in which unidirectional signal flow is effected by using a Hall-effect isolator.

1. The input (operating) frequency is 500 kc. Diode $CR1$ causes an amplified signal to be developed across the primary of transformer $T1$ which is resonated at 500 kc by capacitor $C2$. Transformer $T1$ matches the low input resistance (about 2 ohms) of the Hall-effect isolator to the output resistance of the tunnel diode. An impedance match is achieved when the positive resistance reflected to the primary circuit is slightly less than the negative resistance of diode $CR1$. This condition gives maximum gain with stability. Resistors $R1$ and $R2$ form a voltage divider to bias diode $CR1$. Capacitor $C1$ ac bypasses resistor $R2$ at the operating frequency. Resistors

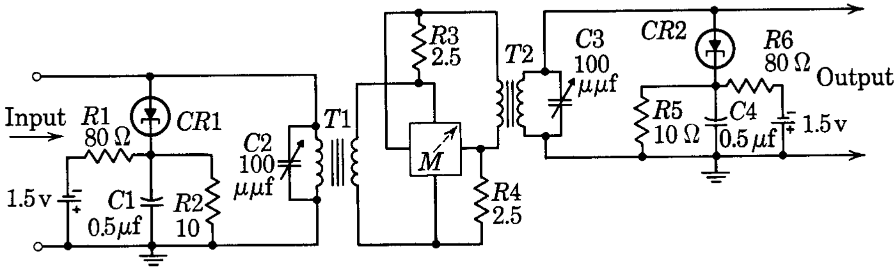


Fig. 5–14. Cascaded tunnel diode amplifiers using Hall-effect isolator and transformer coupling

*R*3 and *R*4 convert the Hall-effect gyrator to Hall-effect isolator by negative feedback.

2. Transformer *T*2 matches the low output impedance of the Hall-effect isolator to the high input impedance of diode *CR*2 and its load circuit. Capacitor *C*3 resonates the secondary of transformer *T*2 to 500 kc. Diode *CR*2 causes an amplified signal to appear across its load resistor (not shown). Resistors *R*5 and *R*6 form a voltage divider to bias diode *CR*2. Capacitor *C*4 ac bypasses resistor *R*5.

b. Matching with Series Resonant Circuits. Figure 5–15 shows two tunnel diode amplifiers coupled through a Hall-effect isolator. Note that in this application only one feedback resistor (*R*3) is used to convert the Hall-effect gyrator to a Hall-effect isolator. The negative feedback current

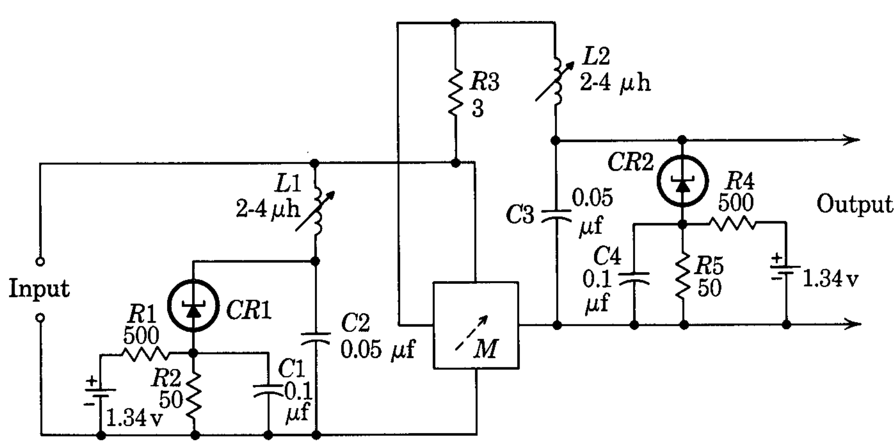


Fig. 5–15. Cascaded tunnel diode amplifiers using Hall-effect isolator with series resonant circuit for impedance matching

through resistor $R3$ is returned to the second stage ($CR2$) through the Hall-effect isolator.

1. In the input circuit, coil $L1$ and capacitor $C2$ form a series circuit resonant at 600 kc. The impedance of the series circuit is low and matches the low input impedance of the Hall-effect isolator. The tunnel diode ($CR1$) *sees* a parallel resonant circuit formed by $L1$ and $C2$ and therefore a high impedance. The circuit is designed so that a virtual impedance match occurs for $CR1$ and maximum gain is achieved. Resistors $R1$ and $R2$ form a voltage divider for biasing $CR1$. Resistor $R2$ which develops the bias voltage is bypassed by capacitor $C1$.
2. Coil $L2$ and capacitor $C3$ form a series resonant (600 kc) circuit and match the low output impedance to the high input impedance formed by tunnel diode $CR2$ and its load (not shown). Resistors $R4$ and $R5$, bypassed by capacitor $C4$, form a voltage divider and establish bias voltage for $CR2$.

5–15. Cascaded Amplifiers Using Skew Isolator

Figure 5–16 shows two tunnel diode amplifiers coupled through a skew isolator. Tunnel diode $CR1$ causes an amplified signal to appear at the input terminals of the skew isolator. Coil $L1$ and capacitor $C2$, resonant at the operating frequency of 1 mc, provide selectivity. Resistor $R3$ is an impedance-matching resistor that reduces the negative resistance of $CR1$ to match the lower input resistance of the skew isolator. Resistors $R1$ and $R2$, bypassed by capacitor $C1$, develop the bias voltage for $CR1$. In the output stage, the negative resistance of tunnel diode $CR2$ is reduced by resistor $R4$ for impedance matching the diode to the load (not shown). Resistors $R5$ and $R6$ form a voltage divider to establish the bias voltage for $CR2$. Resistor $R5$ is ac bypassed by capacitor $C3$. The selectivity of the stage is determined by the parallel resonant circuit formed by coil $L2$ and capacitor $C4$.

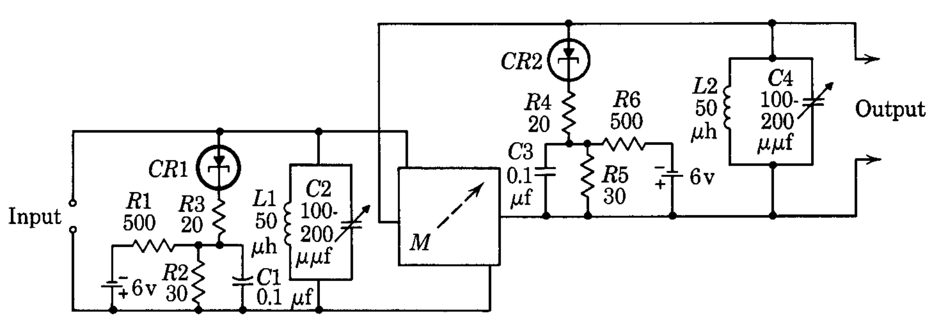


FIG. 5–16. Tunnel diode amplifiers cascaded through a skew isolator

SECTION V. HYBRID JUNCTION

5–16. General

a. A hybrid junction is a four-port, hollow waveguide. It is so constructed that energy at the proper frequency introduced into one port will appear at only two of the remaining three ports. The device can be used to cause a unidirectional flow of energy between cascaded tunnel diode amplifiers. It can be used at frequencies above 300 megacycles. The physical dimensions of hollow waveguides are proportional to the wavelength of the operating frequency. Below 300 mc the hollow waveguide becomes large and cumbersome to use.

b. The principle of operation of the hybrid junction depends upon the propagation of electromagnetic fields. These are reviewed briefly in paragraph 5–17. Cascaded tunnel diode amplifiers using a hybrid junction are discussed in paragraph 5–18.

5–17. Electromagnetic Fields

Electromagnetic fields propagated through hollow waveguides are the same as radio waves propagated through free space. The main difference is that the boundaries of the fields within the waveguide are fixed by the

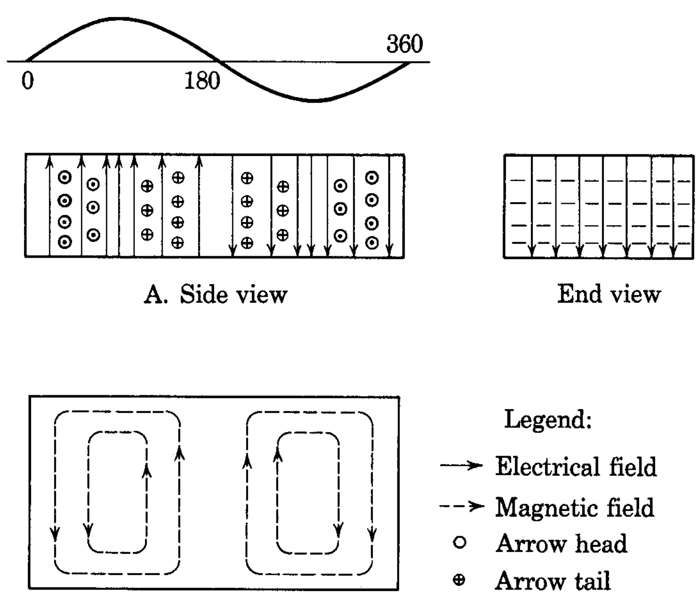


FIG. 5–17. Electromagnetic wave propagated through a hollow waveguide

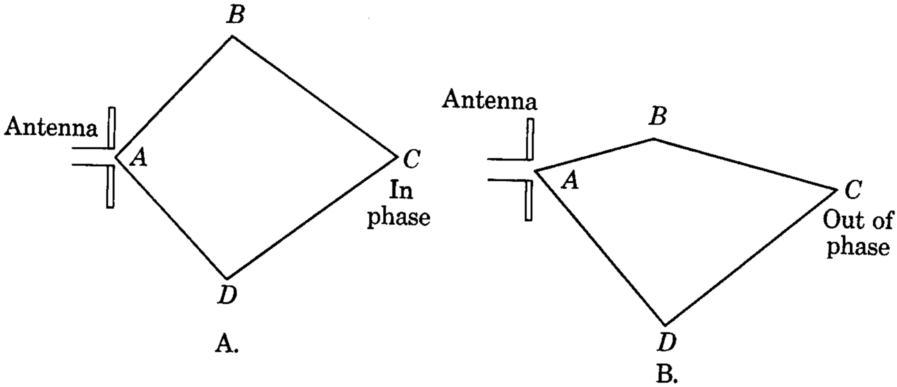


FIG. 5–18. Various paths traveled by portions of the same field

waveguide. The electrical field and the magnetic field are always at right angles to each other. The direction of the electrical field determines the polarity of the field. A vertically or horizontally polarized field means that the plane of the *electrical field* is vertical or horizontal, respectively.

a. Figure 5–17A shows a vertically polarized field within a hollow waveguide. The solid-line arrows represent the electrical field. The intensity of the field is represented by the number of solid-line arrows in a given unit distance. The circle and dot represent a magnetic field emerging from the page; the circle and cross represent a magnetic field entering the page. A top view of the magnetic field is shown in dashed-line arrows in Fig. 5-17B. The intensity and direction of the electrical field is shown also by the sinewave.

b. Figure 5–18A shows an antenna radiating into space. Assume that because of reflections one portion of the radiated energy takes one path (*ABC*) from the antenna to point *C*, and another portion of the energy takes a second path (*ADC*). If the two paths are equal in distance and the fields retain their original polarity, the fields will arrive at point *C* in phase. The intensity of the resultant field at point *C* will be the sum of the two portions of the electrical field. Figure 5–18B also shows an antenna radiation in space. In this case the two portions of radiated energy travel unequal distances to point *C*. If the difference in distance between paths *ABC* and *ADC* equals the distance the signal travels in one cycle of time, the waves will arrive at point *C* 180° out of phase. For instance, a positive peak and a negative peak will arrive at point *C* simultaneously. The fields cancel and the net intensity at point *C* will be zero. Depending on the specific distances, of course, any phase relationships can exist at point *C*. The same phenomena can happen in a hollow waveguide. This principle is employed in the hybrid junction (par. 5–18).

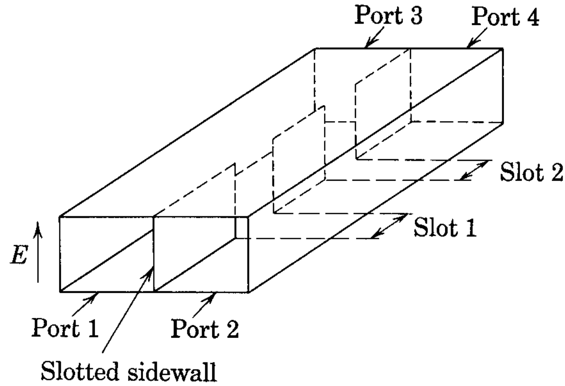


FIG. 5–19. Structure of hybrid junction

5–18. Hybrid Junction

a. The structure of a hybrid junction is shown in Fig. 5–19. It is a hollow waveguide divided into two sections by a sidewall containing two slots. Electrical energy can enter or leave through four openings designated port 1 through port 4. The dimensions of the guide depend upon the wavelength of the operating frequency. The polarity of the electrical field is as indicated by the solid-line arrow marked *E*.

b. If electrical energy is introduced at port 1 (Fig. 5–20A), there will

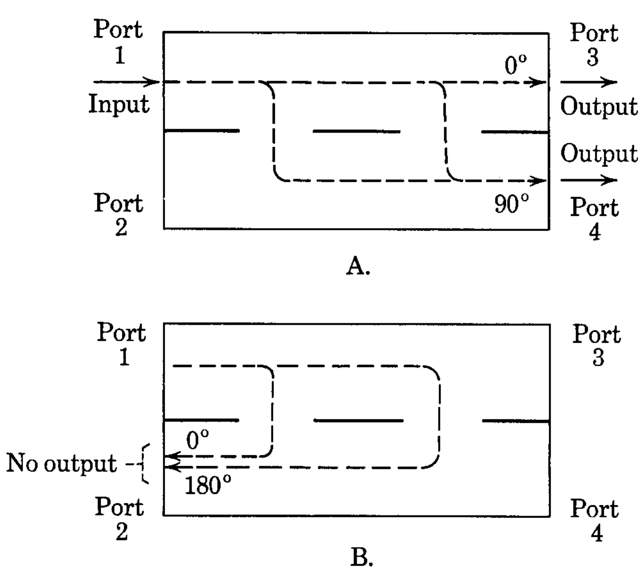


FIG. 5–20. Propagation of fields in hybrid junction

be an output at ports 3 and 4; no output will appear at port 2. Energy will go from port 1 to port 3 directly. Energy will go from port 1 to port 4 through slots 1 and 2; because these two paths through slots 1 and 2 are *equal* in distance, the two portions of the field will be *in phase* at port 4 and permit electrical energy to be transferred to port 4. The 0° phase notation at port 3 indicates that at the instant that a positive peak occurs at port 1, a positive peak also occurs at port 3. Because of the difference in distance traveled by the field in going to ports 3 and 4, the field at port 4 is 90° out of phase with that at port 3.

c. The two portions of the field that enter slots 1 and 2 and travel toward port 2, travel *different* distances. The slots are arranged so that the difference in distances is one wavelength. The two portions of the field arrive at port 2, 180° out of phase, cancel, and prevent transfer of energy from port 1 to port 2.

d. Because of the symmetrical positioning of the slots, energy can be introduced into any port, arrive at the two opposite ports with no energy arriving at the adjacent port. The hybrid junction can also be designed so that *equal amounts* of energy arrive at the output ports.

5–19. Hybrid Coupled Tunnel Diode Amplifiers

Figure 5–21A shows a block diagram of a hybrid junction with a 300-mc signal source connected to port 1, tunnel diode amplifiers connected to ports 3 and 4, and a 50-ohm load connected to port 2.

a. Equal amounts of energy arrive from the signal source at amplifiers No. 1 and No. 2, 0° and 90° out of phase as indicated. The signal is amplified in each amplifier and transferred back to the hybrid junction. The field from amplifier No. 1 to the source is not changed in phase (Fig. 5–21B). The field from amplifier No. 2 to the source is changed in phase *an additional 90°* so that it arrives at the source 180° out of phase with that from amplifier No. 2. These fields cancel, and no energy is transferred to the source from the amplifiers. The field from amplifier No. 1 to the load is shifted in phase by 90°; the field from amplifier No. 2 is not again shifted in phase, but had already experienced a 90° phase shift (Fig. 5–21A). As a result, the two fields from the amplifiers to the load arrive in phase at the load and transfer an amplified signal to the load.

b. A simplified practical circuit of the two-stage amplifier is shown in Fig. 5–22. All connections to the hybrid junction are made through coaxial cables terminated in an inductive loop. The loops act as transformers. The 300-mc signal source at port 1 has an internal resistance of 50 ohms. At port 2 is the 50-ohm load resistor. Each amplifier is as shown at port 3. The input impedance to the amplifier is 50 ohms. Impedance-matching resistor R_Z is greater than the negative resistance of the diode to prevent

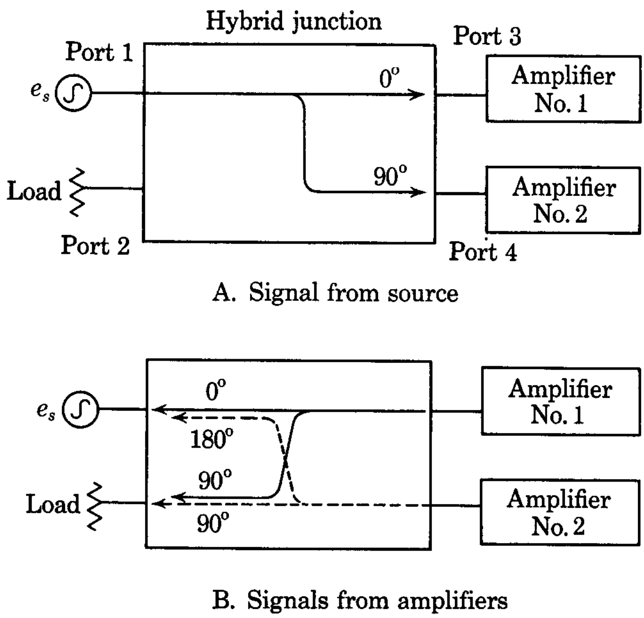


FIG. 5-21. Block diagrams of a two-stage tunnel diode amplifier coupled through a hybrid junction

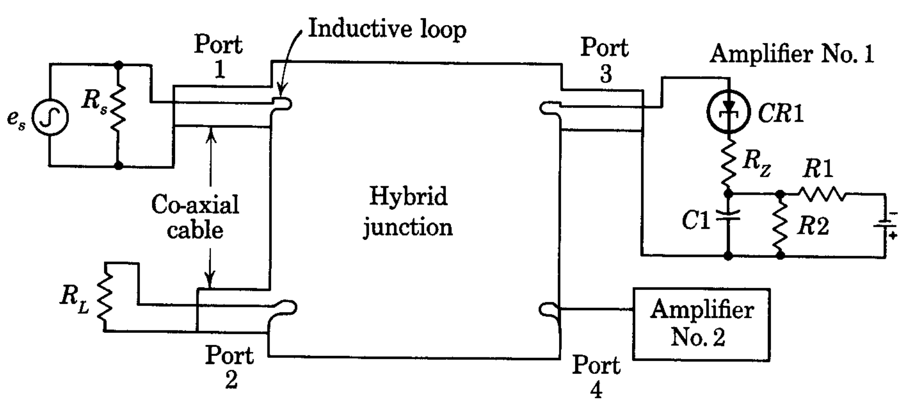


FIG. 5-22. Simplified schematic diagram of cascaded tunnel diode using hybrid junction

oscillation. Resistors $R1$ and $R2$ form a voltage-divider bias network bypassed by capacitor $C1$. A gain of approximately 10 db with a 300-mc (150 – 450 mc) bandwidth was achieved with a low noise figure of 1.7 db.

SECTION VI. FERRITE CIRCULATOR

5–20. General

a. The ferrite circulator embodies a phenomenon first observed by Michael Faraday in 1845. Faraday was trying to prove that light energy is electromagnetic in nature. He did this by passing a linearly polarized wave through a piece of glass that was immersed in a strong magnetic field. The magnetic field was parallel to the direction of the propagation of light. He noticed that the plane of polarization of the light wave was rotated when it passed through the glass. By passing the light in the *opposite* direction he found that the plane of polarization of the light was rotated in the *opposite* direction. The Faraday rotation effect is therefore *nonreciprocal.* It is this fact that makes the phenomenon useful in circulator-isolator devices used today at microwave frequencies. He determined further in his experiments that the amount of rotation is proportional to the strength of the magnetic field and the thickness of the glass.

b. The Faraday rotation effect did not become important in microwave technology, however, until the recent development of ferrites. Ferrites are made from minerals of iron oxide in which some of the iron atoms are replaced with metals, such as nickel, zinc, and manganese. The chemical structure results in a high resistivity (low eddy currents), low-loss, high permeability material. Studies of the Faraday rotation effect, using ferrite materials at microwave frequencies, indicated rotations of the electrical field up to 150° per inch.

c. The Faraday rotation effect depends upon the precession of electrons in a ferrite immersed in a magnetic field (par. 5–21), and the resolution of a linearly polarized field into two rotating fields (par. 5–22).

5–21. Electron Precession

a. An electron spinning naturally about its own axis is shown in Fig. 5–23A. Because the electron has electrical charge, it creates its own magnetic field M. Because it has mass, it also displays an angular mechanical momentum J.

b. When an external magnetic field (H) is applied in the direction shown (Fig. 5–23B), the magnetic field of the electron would normally tend to align itself with the applied field (H). However, the mechanical momentum J acting on the electron, in conjunction with the action of the two

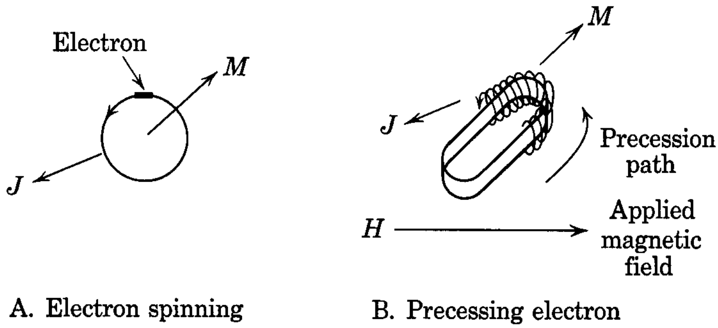


FIG. 5–23. Spinning and precessing electron

magnetic fields, causes the electron to precess in a circular ring as shown. The term *precess* simply means that this circular action takes *precedence* over any other possible action. The net movement of the electron is a spiral about the precession path.

Note: The precession path is shown as a physical ring only to simplify visualizing the process.

c. The natural rate of precession is 2.8×10^6 revolutions per second per oersted of magnetic field. A magnetic field of 4000 oersteds would cause a precession rate of 11.2×10^9 revolutions per second. Actually this system represents a *resonant* system of 11.2 kmc per second. Furthermore, the system will absorb energy from an electrical field rotating in the same direction as that of the precession. An oppositely rotating electrical field will oppose the precession. The amount of energy absorbed from the rotating electrical field depends on the frequency of the electrical field. Maximum energy is absorbed from a similarly rotating electrical field when the frequency of the field is the *same* as the frequency of precession.

5–22. Linearly Polarized Wave

a. A linearly polarized wave is shown in Fig. 5–17A and is discussed in conjunction with hybrid junctions. A simplified representation of the wave is shown in Fig. 5–24. Here the intensity of the field is represented by the *amplitude* of the electrical field vector rather than the proximity of electrical lines. An end view of the electrical vectors (from the direction of propagation) is also shown.

b. Any given vector quantity can be resolved into two separate vector components whose vector sum is the first given quantity. Figure 5–25 shows an oblique view of the linearly polarized wave. Each original electrical field vector (E_1 through E_5) is resolved into two components (E_{R1} and E_{L1} for E_1, E_{R2} and E_{L2} for E_2, etc.). A view of the vectors facing the main

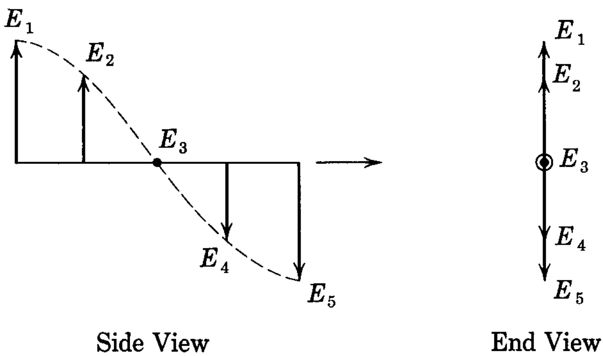


FIG. 5–24. Linearly polarized wave, side and end views (magnetic field not shown)

arrowhead is also shown. As the field moves toward and away from this point, vectors E_{R5}, E_{R4}, E_{R3}, E_{R2}, and E_{R1} would be encountered in that order. These vectors would represent an electrical field rotating in the *counterclockwise direction.* In the same manner vectors E_{L5} through E_{L1} would represent an electrical field rotating in the *clockwise direction.* Thus the linearly polarized wave is shown to consist of two *oppositely* rotating electrical fields. Note, however, that the original field (E_1 through E_5) remains in its original position.

c. Assume that the linearly polarized wave is fed into a hollow waveguide containing a ferrite core at its center (shown subsequently in Fig. 5–27A). With the ferrite immersed in an external magnetic field, the electrons within the ferrite will be precessing at a frequency dependent upon the strength

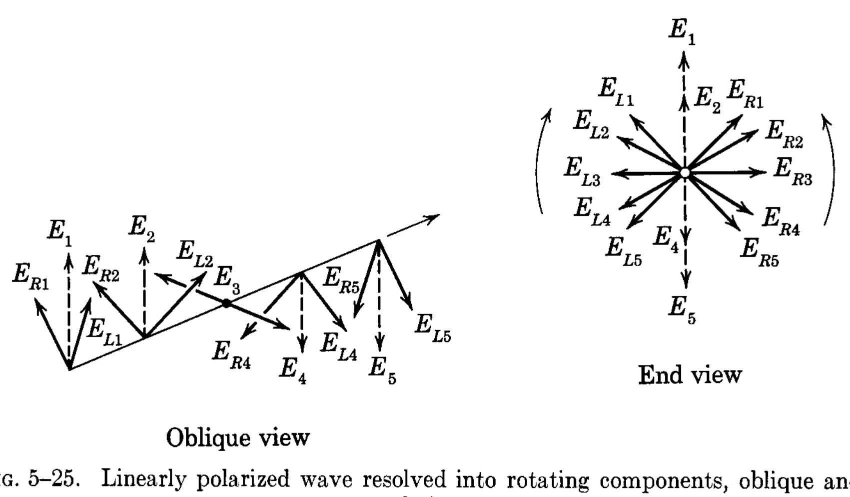


FIG. 5–25. Linearly polarized wave resolved into rotating components, oblique and end view

of the magnetic field. One rotating component (*b* above) of the linearly polarized wave will aid the precessing and be slightly advanced in rotation; the other rotating component will oppose the precessing and be slightly delayed in rotation. An oblique view of the wave *after* it has passed the ferrite is shown in Fig. 5–26. An end view at the arrowhead is also shown. The net field (E_1 through E_5) is the vector sum of the components. Note that the net field (still linearly polarized) is rotated with respect to the

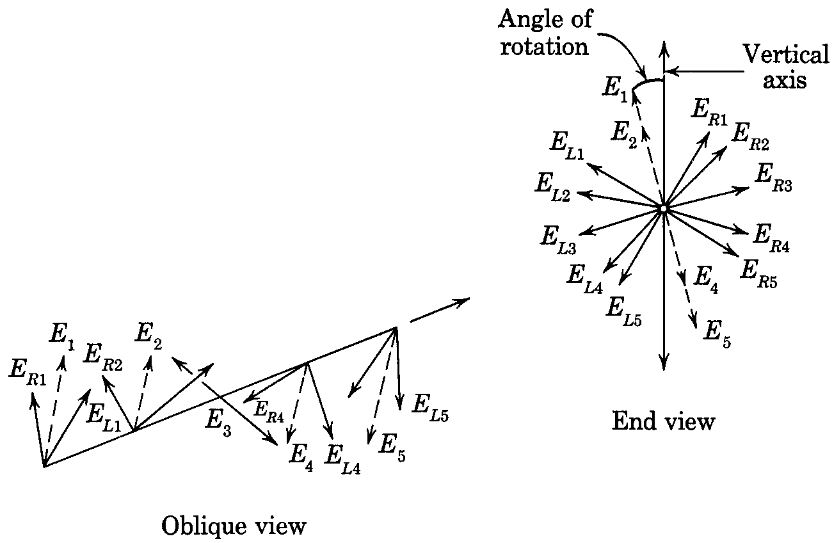


FIG. 5–26. Oblique and end view of linearly polarized wave after passing through magnetized ferrite

position it would have occupied had the magnetized ferrite not been present. This rotation of the linearly polarized field is used in the ferrite circulator device (par. 5–23).

5–23. Ferrite Isolator

a. The ferrite isolator is a two-port device that permits energy flow in one direction only. The basic structure of a ferrite isolator is shown in Fig. 5–27A. A longitudinal magnetic field is supplied by a ring-type permanent magnet mounted on the exterior of a hollow circular waveguide. In the center of the waveguide is a ferrite core supported by low loss dielectric material. In passing through the isolator a linearly polarized wave is rotated 45°. The amount of rotation is fixed by the strength of the magnetic field and the length of the ferrite core.

b. Figure 5–27B shows a vertically polarized wave entering port 1 of the isolator and leaving port 2 rotated 45°. Figure 5–27C shows an oblique view of the vertically polarized wave with the field rotated 45° and emerging from port 2. If a pick-up device (loop acting as transformer) is oriented at port 2 in the same direction as the emerging field, the energy can be transferred to a load, or amplifier placed at port 2.

c. If a portion of the original signal or an amplified signal (dashed-line arrow, Fig. 5–27C) is reflected back to port 1, it will also be rotated 45° but in the opposite direction. It will arrive at port 1 horizontally polarized. Since the pick-up device normally placed at port 1 will absorb energy only

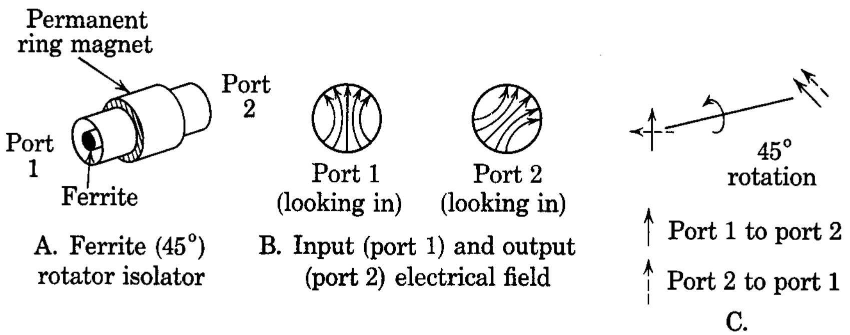


FIG. 5–27. Basic structure and operation of 45° rotator isolator

from vertically polarized waves, no energy will be transferred from port 2 to port 1. Port 1 therefore is effectively isolated from port 2.

5–24. Ferrite Circulator

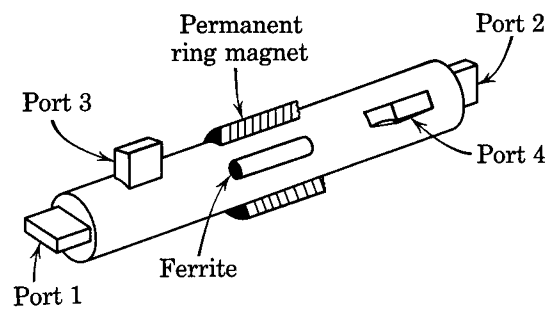
a. The basic structure of a four-port ferrite circulator (45° rotator type) is shown in Fig. 5–28A. The structure of the ferrite core and permanent ring-type magnet is identical to that of the ferrite isolator (par. 5–23). The ports are positioned so that, with the 45° rotation provided, energy flows from port 1 to port 2 to port 3 to port 4, and back to port 1.

b. To understand this energy flow, refer to Fig. 5–28B. The orientation of the pick-up and transmitting devices at the four ports is indicated by heavy lines (1 through 4) placed in the *X*-*Y*-*Z* coordinates. These devices operate in the same manner as *dipole* antennas. Each dipole is in a plane parallel to the *Y*-*Z* plane. Port 1 *dipole* is parallel to the *Y* axis; it is vertical. Port 3 dipole is horizontal. Port 4 dipole intersects the *X*-*Z* plane at 45° as shown. Port 2 dipole intersects the *X*-*Z* plane at a 45° angle as shown.

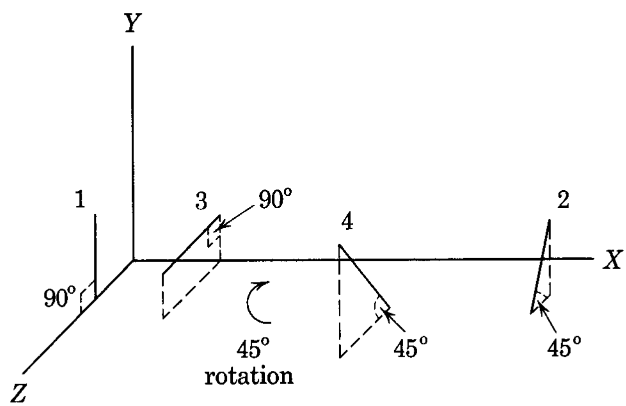
c. Assume a vertically polarized signal is introduced at port 1.

1. The signal passes port 3 undisturbed (at right angles to port 3 dipole), is rotated 45° by the ferrite, passes port 4 undisturbed, and is picked up by port 2 *dipole* with which it is in phase.

2. Energy reflected or amplified at port 2 is polarized at a 45° angle as



A. Ferrite (45°) rotator circulator



B. Orientation of pick-up and transmitting devices at ports

FIG. 5–28. Four-port ferrite circulator and orientation of dipoles

indicated. The energy passes port 4 undisturbed, is rotated 45° by the ferrite, and is picked up by port 3 dipole with which it is in phase. Port 1 *cannot* pick up energy from port 2.

3. Energy reflected or amplified at port 3 is horizontally polarized and cannot be picked up by port 1. The signal is rotated 45° by the ferrite and picked up by port 4 with which it is in phase. Port 2 cannot pick up energy from port 3.

4. Energy reflected or amplified at port 4 is polarized in the same direction as port 4 dipole is positioned. Port 4 energy cannot be picked up by port 2. The signal is rotated 45° by the ferrite, undisturbed by port 3, and picked up by port 1 with which it is in phase.

d. When used as an isolator circulator, port 4 is usually terminated in a matched load. Energy arriving at port 4 will then be absorbed from port 3 and *not* transmitted to port 1. When so used, the device is represented graphically as shown in Fig. 5–29. Numbers 1, 2, 3 represent ports, 1, 2, 3, respectively. Port 4 (not numbered) is shown terminated in a matched resistor. The arrow indicates the direction of energy flow, namely from port 1, to 2, to 3. One application of this device for cascading tunnel diode amplifiers is discussed in paragraph 5–25.

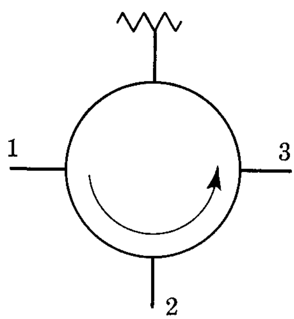


Fig. 5–29. Graphical representation of four-port ferrite circulator, terminated in matched load at port 4

5–25. Cascaded Amplifiers Using Ferrite Circulator

a. A 4-kmc tunnel diode (*CR*1) amplifier is shown in Fig. 5–30A. Electrical energy is conducted by means of hollow waveguides, which are most efficient in the microwave region. A ferrite circulator (par. 5–24) directs the flow of energy from the signal source to the tunnel diode and to port 3. Energy does not flow in the reverse direction. Port 4 is terminated in a matched load to absorb any energy that reaches it from port 3. Although the energy that reaches port 4 represents a loss, the gain of the amplifier is 20 db with a bandwidth of 20 mc and a noise figure of 6 db. Tunnel diode *CR*1 properly biased is placed in a single port cavity resonator that acts like a tank circuit at microwave frequencies. The power output of the stage is one microwatt. The dielectric plug absorbs energy in the same manner as a resistor in wired (conventional) circuits. Enough energy is absorbed by the dielectric plug to prevent oscillation of the stage.

b. Three such similar stages are shown cascaded in Fig. 5–30B.

5–26. Summary

a. Each set of terminals associated with a device is called a *port.* Thus a junction transistor is considered a two-port device. A tunnel diode is considered a one-port device.

b. Any opening in a hollow waveguide or cavity resonator which can be used to introduce or extract energy is called a *port.*

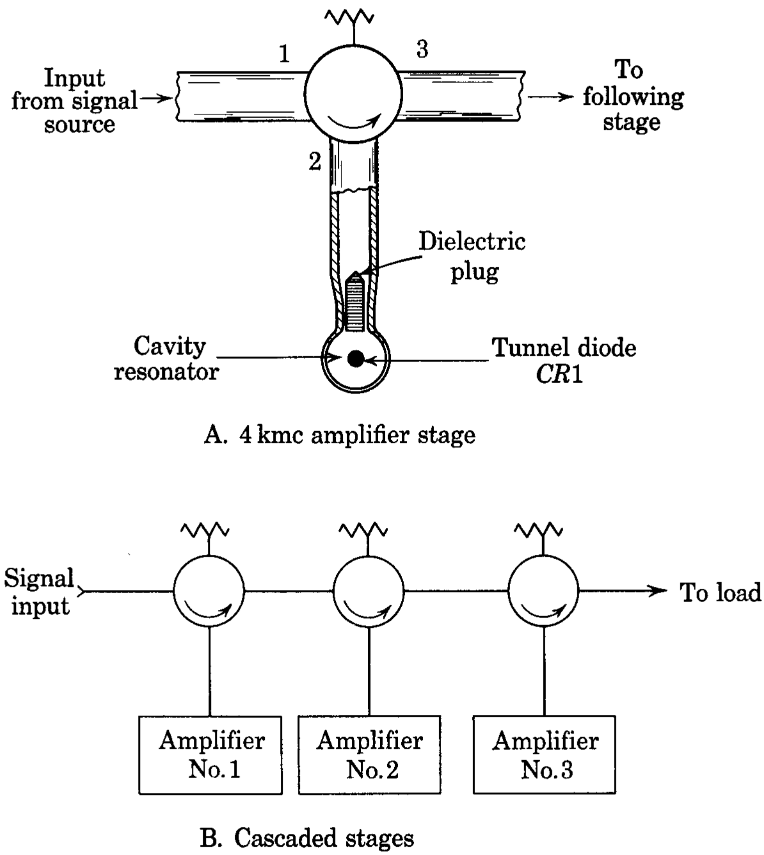


FIG. 5-30. Tunnel diode (4 kmc) amplifier stage using ferrite circulator, and three cascaded stages

c. The tunnel diode, a one-port device, is a bilateral active unit.

d. To cascade tunnel diode amplifiers, the output of a succeeding stage must be isolated from the output-input of a previous stage; i.e., a unidirectional (unilateral) flow of energy must be assured.

e. A unidirectional flow of energy between cascaded tunnel diodes can be achieved by using:

1. The matched transmission line technique
2. The matched (double-open end) quarter-wave line technique
3. The Hall-effect or skew isolator
4. The hybrid junction
5. The ferrite isolator or circulator

f. If a signal source is matched to the characteristic impedance of a transmission line which is terminated in a resistance equal to the characteristic impedance, reflections of energy from the load to the source will not occur. By cascading series-parallel arrangements of tunnel diodes (negative resistances), a constant impedance match similar to a matched transmission line can be achieved. This condition results in a unidirectional flow of energy.

g. The double-open end quarter-wave line displays a finite positive resistance at each end. The simulated open at each end can be obtained by a parallel arrangement of equal values of positive resistance and negative (tunnel diode) resistance. By proper matching, unidirectional flow of energy will result.

h. The Hall-effect gyrator, a nonreciprocal device, operates on the principle of the rotation of equipotential lines of force in a material conducting electricity and exposed to an external magnetic field.

i. The Hall-effect gyrator can be used as an isolator by adding one or two negative feedback (shunt) resistors to the circuit.

j. The skew isolator operates on the principle of the Hall effect. Isolation between circuits is obtained by an asymmetrical positioning of one pair of terminals.

k. The hybrid junction is a four-port device that depends for its operation on the difference in path distances traveled by electromagnetic waves in a hollow waveguide. It is so designed that electromagnetic waves introduced at one port will appear at two oppostie ports 90° out of phase with each other; no energy will appear at the adjacent port.

l. The ferrite isolator and circulator operate on the Faraday rotation effect which depends upon:

1. Precession of the electrons of the ferrite when immersed in an external magnetic field.
2. The resolution of a propagated linearly polarized electromagnetic field into two oppositely rotating components.

Chapter 6

SEMICONDUCTOR COMBINATIONS AND CHARACTERISTICS

SECTION I. GENERAL

6–1. Tunnel Diode Combinations

Tunnel diodes can be combined in series or parallel arrangements to effect characteristics and properties which suggest numerous practical applications. The development of the characteristics for several series arrangements is covered in paragraphs 6–4, 6–5, and 6–6. Characteristics of parallel and other arrangements are discussed in paragraphs 6–7 through 6–12. Practical applications of some of these combinations are covered in subsequent chapters.

6–2. Tunnel Diode and Backward Diode

The backward diode consists of a p_T-n junction that, for a given value of voltage, conducts current more heavily in the reverse-biased direction than in the forward-biased direction. It is a by-product of extensive research conducted on tunnel diodes. The backward diode can be used in circuits by itself as a nonlinear device, or in combinations with the tunnel diode to achieve certain desirable characteristics. The properties of the backward diode are discussed in paragraph 6–13. Combined characteristics of tunnel and backward diodes are discussed in paragraph 6–14.

6–3. Transistors and Tunnel Diodes

A natural combination of semiconductor devices is the widely used junction transistor and the tunnel diode. Numerous combinations are possible to obtain certain desirable effects. However, this chapter covers only those combinations formed especially to change the characteristics of the tunnel diode. For instance, the operating current level of the tunnel diode can be increased (par. 6–15) or decreased (par. 6–16). In addition its voltage-current characteristic can be changed from an S-type negative resistance

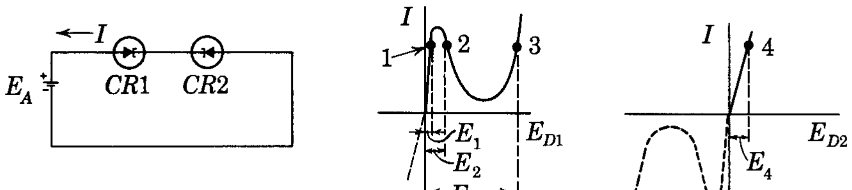
characteristic to an N-type negative resistance characteristic (par. 6–18). A brief discussion of S- and N-type characteristics is given in paragraph 6–17. A brief review of the basic theory of operation of transistors is given in Appendix C.

SECTION II. SERIES AND PARALLEL TUNNEL DIODE COMBINATIONS

6–4. Back-to-Back Tunnel Diodes

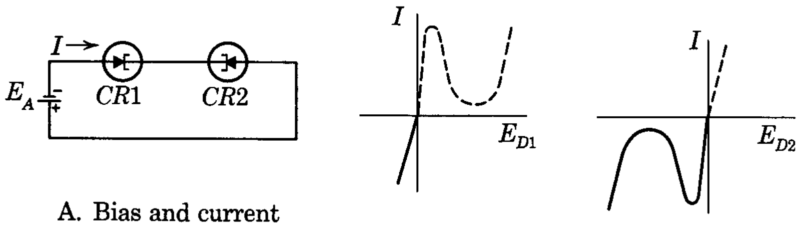
Two tunnel diodes can be placed in series and back-to-back (shown subsequently in Fig. 6–3A) to achieve the overall symmetrical characteristic shown in Fig. 6–3B. Such an arrangement can be used as a symmetrical switching device. The current and voltage relationships involved to produce the characteristic are discussed in *a* and *b* below.

a. Figure 6–1A shows the circuit arrangement with an applied voltage (E_A) having the indicated polarity. Diode $CR1$ is forward biased and diode $CR2$ is reverse biased. The active portions of the current-voltage chart of diodes $CR1$ and $CR2$ for the indicated polarity are shown as solid lines in B and C, respectively, Fig. 6–1. The inactive portions are shown as dashed lines. Diode $CR2$ chart is shown inverted because the current through diode $CR2$ is the same as that through diode $CR1$ which is active in the first quadrant. The points marked 1, 2, 3, and 4 are arbitrarily chosen and represent the *same* current magnitude. If points 1 and 4 are the operating points, then the applied voltage (E_A) must equal the voltage drop (E_1) across diode $CR1$ and the voltage drop (E_4) across diode $CR2$. If points 2 and 4 are the operating points, then $E_A = E_2 + E_4$; if points 3 and 4 are the operating points, then $E_A = E_3 + E_4$. The latter equations indicate that the composite curve for the two diodes would have the same current magnitude as one forward-biased diode; but the voltage at each point would be offset (increased) by the small voltage across the reverse-biased diode. This portion of the composite curve is shown as the upper half of the chart (Fig. 6–3B).



A. Back-to-back tunnel diodes B. $CR1$ curve C. $CR2$ curve

FIG. 6–1. Back-to-back tunnel diodes with positive voltage applied



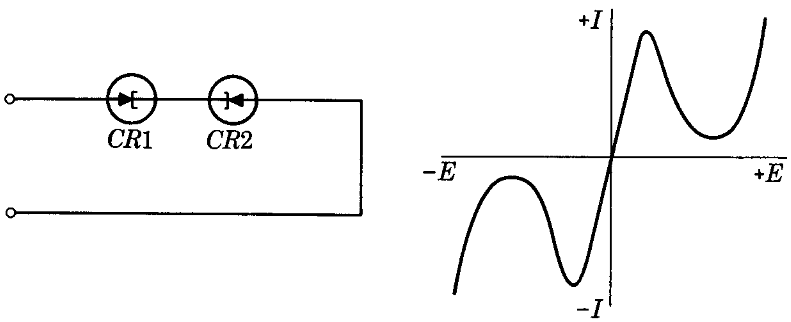
A. Bias and current reversed B. *CR*1 curve C. *CR*2 curve

FIG. 6–2. Back-to-back tunnel diodes with negative voltage applied

b. Figure 6–2A shows the circuit arrangement with the voltage applied in the opposite direction. Diode *CR*1 is now reverse biased (solid-line portion of chart Fig. 6–2B). As in the previous case (*a* above), the composite curve would take the approximate shape of the diode *CR*2 chart with voltage offset (increased negatively) by the small voltage drop across diode *CR*1. This portion of the composite curve is shown as the lower half of the chart (Fig. 6–3B). If the two diodes are matched, the composite chart will be symmetrical and display equal negative-resistance characteristics with positive or negative voltage applications.

6–5. Diodes in Series, Increasing Peak Currents

Figure 6–4A shows three tunnel diodes placed in series in such a manner that an applied voltage will forward bias or reverse bias all three simultaneously. The diodes are selected so that diode *CR*2 has a higher peak current than diode *CR*1; diode *CR*3 has a higher peak current than diode *CR*2. The resultant composite characteristic for this arrangement is shown in Fig. 6–4B. The characteristic indicates use of this arrangement as a four-state switch. Such an application is discussed in Chapter 8. Any



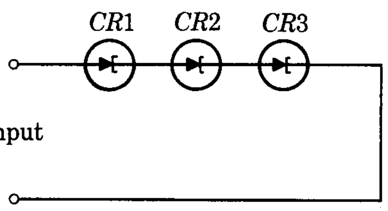
A. Back-to-back tunnel diodes B. Composite characteristic

FIG. 6–3. Composite symmetrical characteristic of back-to-back tunnel diodes

number of diodes can be used in series to produce the desired number of states. This paragraph discusses the generation of the composite characteristic.

a. First State. Generation of the characteristic is best explained by connecting the three diodes in a circuit (Fig. 6–5A). Bias is supplied from a *constant current* source consisting of battery E_B and resistor R_s; i.e., the current flow (I_s) is determined mainly by source resistor R_s, whereas the applied voltage (E_A) is determined by the resistance of the load (the three diodes). In addition, coupling capacitor $C1$ permits injection of pulses to the diodes; for the first state the capacitor and pulse are not required. Figure 6–5B shows the current-voltage characteristic of each diode and the voltage drop across each. The current through each is the same. Note that each diode is biased in the low-voltage state. The data shown on the individual charts can be transferred to the composite chart (Fig. 6–5C). The corresponding voltage drops are shown. The third and fifth dashed lines correspond to the injection currents of diodes $CR1$ and $CR2$, respectively. The dashed lines also correspond to the zero points of diodes $CR2$ and $CR3$. This condition of the circuit arrangement is referred to as the first state (of the four-state switch).

b. Second State. Assume that a short-duration pulse of current is introduced into the circuit through capacitor $C1$. Assume also that the resultant total current equals or just exceeds the peak current of diode $CR1$ only. When the peak current of diode $CR1$ is reached, the diode will switch rapidly to its high-voltage state (dashed line, Fig. 6–6A). This action occurs so rapidly that the contour of the diode $CR1$ characteristic is not followed in switching; the region between peak and valley current is highly unstable when the diode is fed from a current source. The effect of the pulse on diodes $CR2$ and $CR3$ is also shown by dashed lines. Note that diodes $CR2$ and $CR3$ remain in the low-voltage state. This condition can be transferred to the composite characteristic (Fig. 6–6B) and is referred to as the second state (of the four-state switch).



A. Three series diodes

B. Composite characteristic

FIG. 6–4. Three tunnel diodes in series and resultant composite characteristic

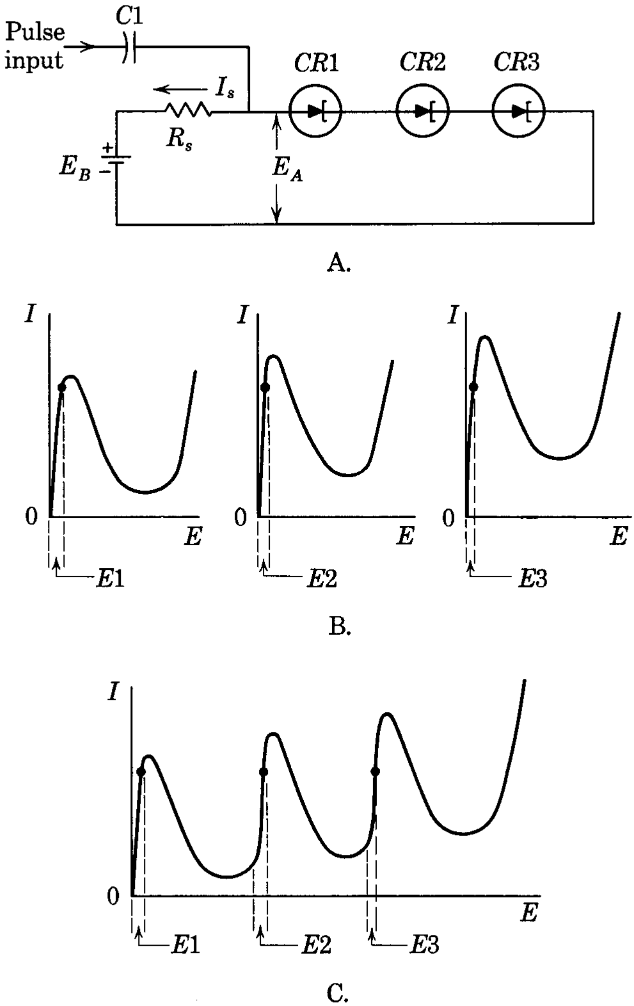


FIG. 6–5. Four-state switch showing individual and composite characteristics for first state

c. Third and Fourth States. Assume that a second short-duration current pulse is introduced through capacitor $C1$ so that the total current equals or just exceeds the peak current of diode $CR2$. Diode $CR2$ will now switch to its high-voltage state (dashed lines, Fig. 6–7A). Diode $CR1$ is in its high-voltage state (*b* above). Only diode $CR3$ remains in its low-voltage state. The fourth state is reached when a third pulse is introduced to switch diode $CR3$ to its high-voltage state (Fig. 6–7B). Each diode is in its high-voltage state (Fig. 6–7C). Note that each state raises the output voltage

level across the three diodes. To return to the first state (*a* above) the diodes must be momentarily shorted, or the bias circuit opened.

6-6. Constant-Voltage Source

a. A tunnel diode in parallel with a load resistor and fed from a constant-current source will provide a constant-*voltage* source (Fig. 6-8A). The constant-current source consists of battery E_B and series resistor R_s. The source current (I_s) is determined mainly by resistor R_s and is calculated by dividing E_B by R_s.

b. Figure 6-8B shows the current-voltage chart with the source current (I_s) as a dashed horizontal (constant) line; the load line (R_{L1}) is also shown. For clarity, the effect of the tunnel diode ($CR1$) is omitted. The vertical distance between the load line and the dashed line represents the current flow through resistor R_{L1} for given voltage. The load line is determined as follows:

1. Assume all the current (I_s) flows through resistor R_{L1}. Multiply I_s by the R_{L1} resistance value to obtain the point of intersection on the voltage axis.

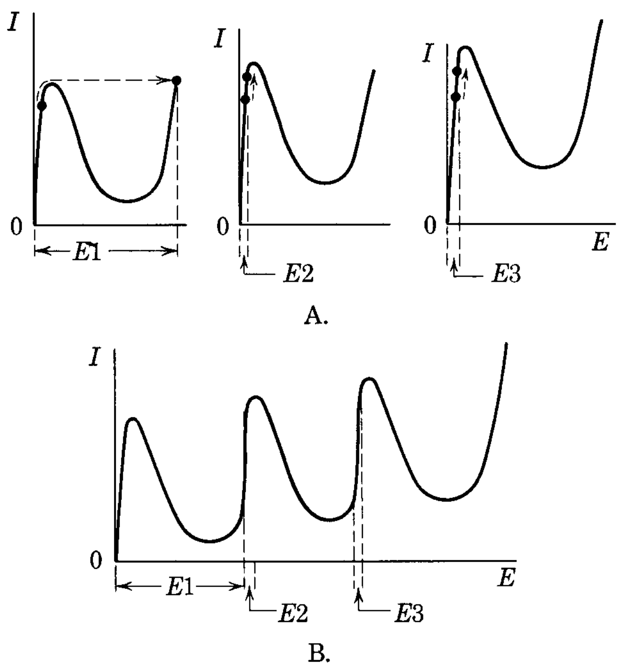


FIG. 6-6. Individual and composite characteristics of four-state switch for second state

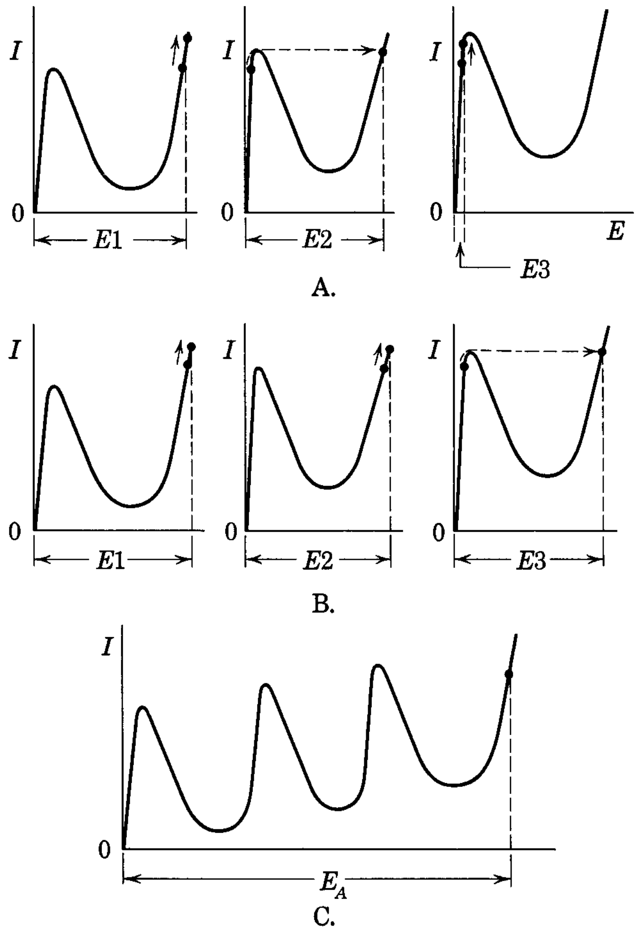


FIG. 6–7. Individual and composite characteristics of four-state switch for third and fourth states

2. The point of intersection of the load line on the current axis is I_s. This condition assumes *no* current flow through R_{L1} and therefore the voltage drop across it is zero. This line is also designated I_s–I_L since each point on the line (read on the current axis) represents the total current (I_s) less the current through R_{L1}.

c. Figure 6–8C represents the same current-voltage chart (*b* above) with the tunnel diode characteristic added. The only point of operation for the circuit is the point of intersection of the load line (R_{L1}) and the tunnel diode characteristic. Load resistance can be increased as indicated by

R_{L2}, R_{L3}, etc., to an infinite value (open load). The current drawn by the various load resistors will vary from a high value (intersection of R_{L1} and tunnel diode characteristic) to zero value (open load).

d. Figure 6–8D shows the total load resistance range, load current (I_L) range, the load voltage (E_L) range. Note that the voltage range is very narrow (almost constant) for a wide variation in load current. This circuit can be used to bias other tunnel diode circuits, such as oscillators, amplifiers, mixers, and demodulators.

6–7. Coupled-Pair or Twin-Diode Arrangement

The coupled-pair or twin-diode arrangement (Fig. 6–9A) is used extensively in multivibrator and pulse and switching circuits (Chs. 7 and 8). In all applications, a circuit element is connected to terminal 2–2′ so that there is an interaction or *coupling* between the two halves; therefore the term *coupled pair.* Also, diodes *CR*1 and *CR*2 are identical diodes; thus the alternate designation *twin diodes.* Analysis of the coupled pair indicates that the arrangement is best considered a *single device* consisting of two ports; terminals 1–1′ constitute one port and terminals 2–2′ constitute an-

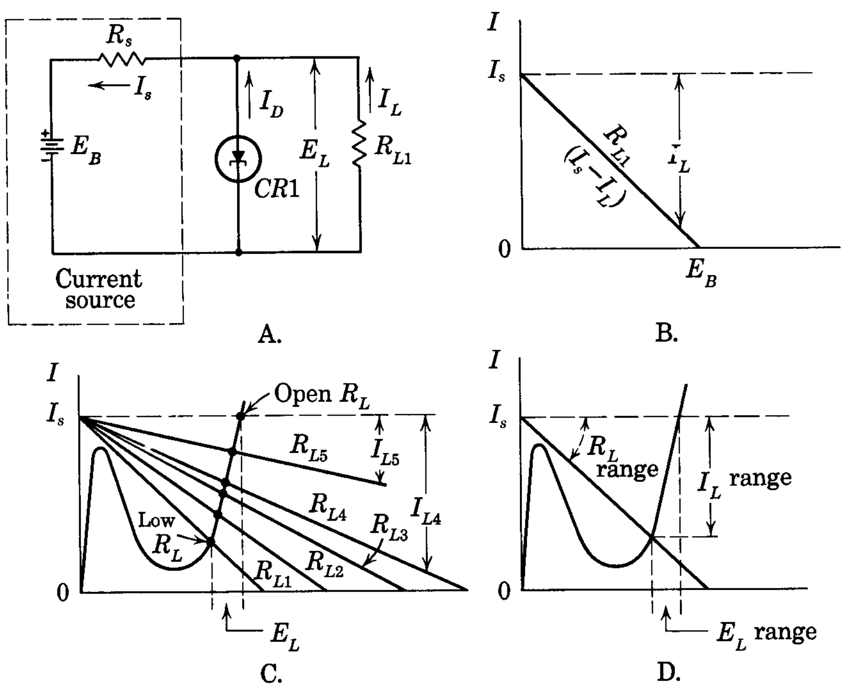


FIG. 6–8. Tunnel diode circuit (and current-voltage characteristics) used to produce constant voltage source

other port. The most important aspect of this circuit is the current-voltage relationship that exists at terminals 2–2′. The analysis in *a* below is based upon a fixed bias voltage $(E_a + E_b)$ at terminals 1–1′ while the supply voltage at terminals 2–2′ is varied. In specific applications the voltage at

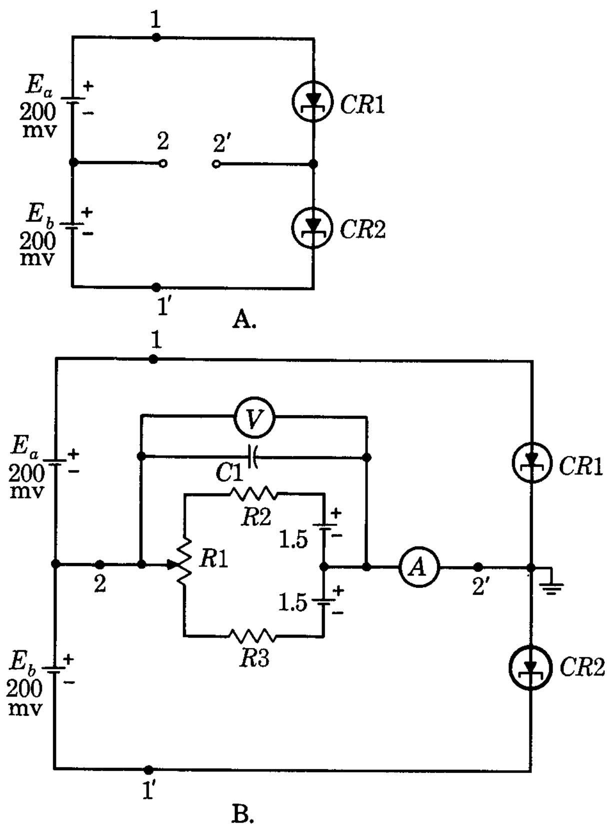


FIG. 6–9. Two-port coupled-pair arrangement and simplified circuit used to obtain current-voltage characteristics

terminals 1–1′ could be a sinewave or even a pulsed signal. The effect of variations in bias voltage $(E_a + E_b)$ on the current-voltage characteristic is discussed in *b* below.

Note: Throughout this discussion voltage E_a equals voltage E_b.

a. Figure 6–9B is a simplified version of a circuit arrangement used to determine the current-voltage curve. A power supply bias has been added; it is capable of varying the voltage at terminal 2 from −200 mv to +200 mv with respect to terminal 2′. Variable resistor $R1$ is used to vary the volt-

age; when $R1$ is at midrange, the applied voltage is zero. Resistors $R2$ and $R3$ are equal in value and limit the voltage swing to the desired value. Voltmeter V measures the applied voltage and ammeter A measures the current. Capacitor $C1$ bypasses ac and surge voltages around the voltmeter.

1. When the voltage at terminals 2–2′ is varied in steps from −200 mv to +200 mv, the current-voltage characteristic at terminals 2–2′ (Fig. 6–10) is obtained. Note that the characteristic has two positive-resistance regions and one negative-resistance region, as is the case with a single

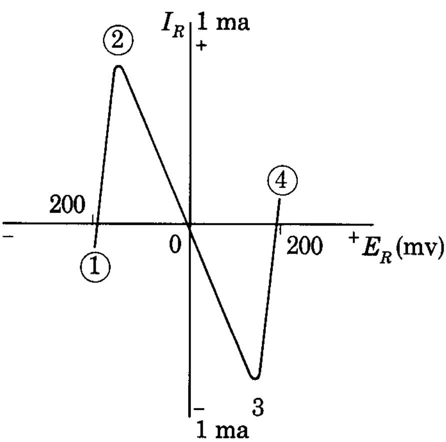


FIG. 6–10. Coupled-pair current-voltage characteristic at terminals 2–2′

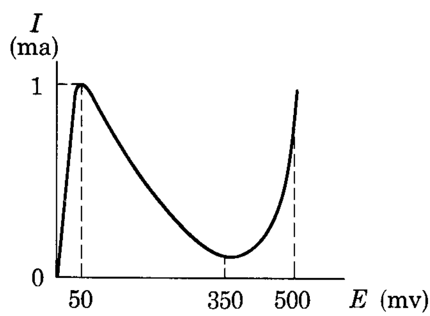


FIG. 6–11. Single-diode ($CR1$ or $CR2$) current-voltage characteristic

tunnel diode. From point 1 to point 2 the resistance is positive because the current is *positive going* and the voltage is positive going (less negative). From point 2 to zero the resistance is negative because the current is *negative going* (less positive) and the voltage is *positive going;* from zero to point 3 the resistance is still negative because the current is *negative going* and the voltage is *positive going.* From point 3 to point 4 the resistance is positive because the current is *positive going* (less negative) and the voltage is *positive going.* Having two positive-resistance regions and one negative-resistance region suggests that the coupled pair can be used in any circuit, in which a single tunnel diode can be used. Some of the advantages of the coupled pair over a single diode are discussed in *c* below.

2. The curve obtained (1 above) by using a practical circuit (Fig. 6–9B) can also be obtained by using graphs. Figure 6–11 shows the current-voltage chart of diode $CR1$ or diode $CR2$. Note that peak current occurs at

50 mv, valley current at 350 mv, and forward injection current at 500 mv. Let it be assumed that all of the current of diode $CR1$ passes through terminals 2–2′. In that circumstance the diode characteristic must be superimposed on the voltage axis representing terminals 2–2′. This condition is shown in Fig. 6–12A. Note that when the voltage at terminal 2 is -200 mv with respect to ground, diode $CR1$ current is zero; the supply voltage is equal and opposite to the bias voltage (E_a). As the negative potential at terminal 2 is reduced, diode $CR1$ conducts and reaches its peak current when terminal 2 potential is -150 mv; the net forward bias on diode $CR1$ is $+50$ mv. Other points on the curve are obtained by considering the net effect of bias and supply voltage on diode $CR1$. In the same manner let it be assumed that all of diode $CR2$ current passes through terminals 2–2′. The characteristic of this diode must also be drawn on the voltage axis of terminals 2–2′ (Fig. 6–12B). In this case when the potential at terminal 2 is $+200$ mv, diode $CR2$ current is zero; as the supply voltage is reduced, diode $CR2$ conducts and reaches its peak at $+150$ mv. Note that when the supply voltage reaches -150 mv, diode $CR2$ is at its valley current point. This condition with diode $CR1$ conducting at peak and diode $CR2$ at valley is shown in Fig. 6–13B. The actual current flow through terminals 2–2′ is the *difference* of the diode currents because these currents move in opposite directions through terminals 2–2′. To obtain the actual current flow then, one of the curves (Fig. 6–12) must be inverted with respect to the other and the sum of the two taken. The inversion and summation is shown in Fig. 6–13A. The resultant current-voltage chart for terminals 2–2′ is shown in dashed lines and is identical with that (Fig. 6–10) obtained with the practical circuit.

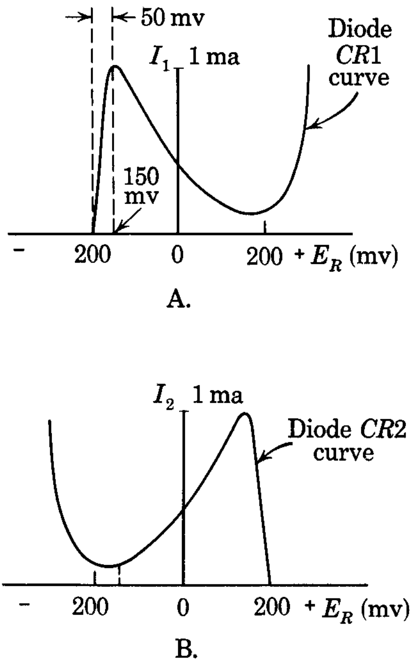


FIG. 6–12. Diode $CR1$ and diode $CR2$ current-voltage characteristic superimposed on terminals 2–2′ voltage axis

b. The effect of changing the bias voltage (E_a and E_b) is shown in Fig. 6–14.

1. In the example in *a* above, $E_a + E_b = 400$ mv. This total applied bias voltage equals the peak voltage (50 mv) of the diode plus the valley voltage (350 mv). In this case it was shown that the peak current of one diode occurs simultaneously with the valley (or minimum) current of the other. It can be expected therefore that when $E_a + E_b$ equals the peak voltage plus the valley voltage of the particular type diode, the maximum peak current will occur at terminals 2–2′, since this current is the difference of the two. This is the actual case and can be shown by further graphical

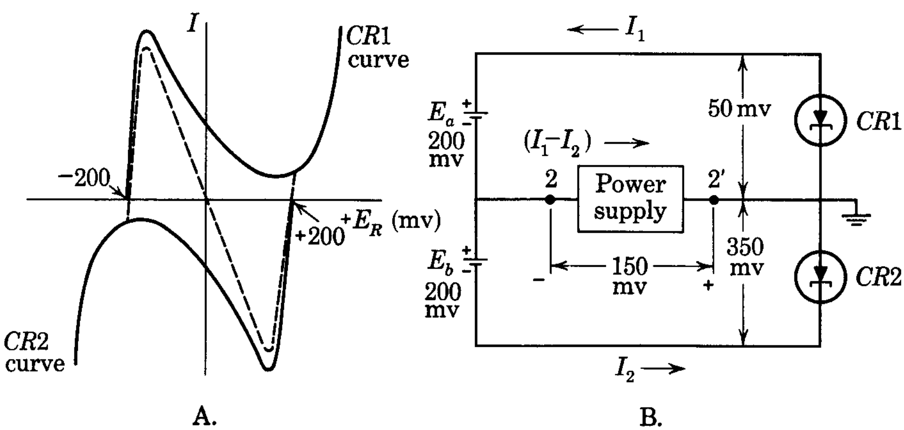


FIG. 6–13. Diode *CR*2 curve inverted to obtain coupled-pair characteristic, and circuit condition when diode *CR*1 draws peak current

analysis or by mathematical analysis beyond the scope of this book. The effect of an increase or a decrease in $E_a + E_b$ is discussed in 2 and 3 below, respectively.

2. With E_a and E_b increased to 280 mv each (560 mv total), the inversion of one diode curve and summation of the two curves with the resultant (dash line) curve is shown in Fig. 6–14A. The resultant curve is obtained with the same graphical technique used in *a* above. In this case, however, diode *CR*1 current is *zero* when terminal 2 potential is negative and equal to E_a (280 mv); diode *CR*1 current is zero when terminal 2 potential is positive and equal to E_b (280 mv). The graphical procedure otherwise is the same. Note that with the increased bias voltage the peak current of the resultant curve is reduced, as might have been expected (1 above), and the voltage at which the peak current occurs is also reduced. It can be shown by mathematical or further graphical analysis that *when the bias voltage equals the diode valley voltage* (i.e., $E_a = E_b = E_v$) *or is greater, no negative-resistance region occurs.* In most applications, therefore, the total bias voltage ($E_a + E_b$) is kept below *twice* the diode valley voltage.

3. With E_a and E_b *decreased* to 50 mv each (100 mv total), the inversion of one diode curve and summation of the two curves with the resultant (dash line) curve is shown in Fig. 6–14B. In this case diode $CR1$ current is *zero*

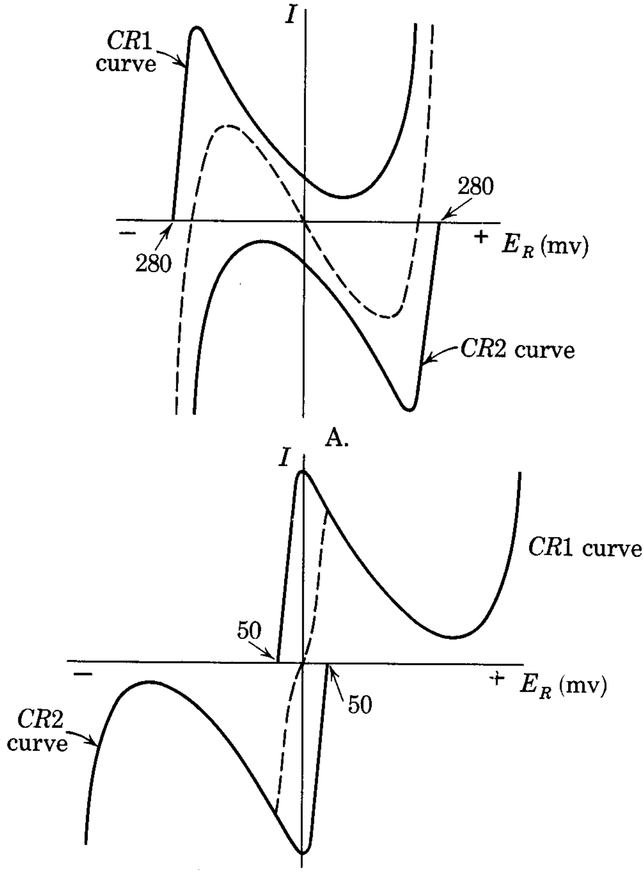


FIG. 6–14. Coupled-pair characteristics with increased and decreased bias voltage

when terminal 2 potential is negative and equal to E_a (50 mv); diode $CR2$ current is zero when terminal 2 potential is positive and equal to E_b (50 mv). Note that no negative-resistance region occurs in the resultant curve. *When the bias voltage equals the diode peak voltage* (i.e., $E_a = E_b = E_p$) *or is less, no negative-resistance region occurs.* In most applications, therefore, the total bias voltage $(E_a + E_b)$ is made larger than *twice* the diode peak voltage.

c. If the current-voltage curves obtained in *a* and *b* above are placed together (Fig. 6–15), a *family* of curves is obtained to describe the coupled pair now being considered a two-port device. This *family* of curves implies one of the main advantages of the coupled pair over the single diode; namely, that *the current-voltage characteristic can be changed by changing the bias voltage.* Note that, unlike the single diode, *the slopes of the coupled-pair characteristic in the positive-resistance regions are the same.* Therefore the resistances of the coupled pair in these regions are equal. The importance of this fact is covered in Chapter 7.

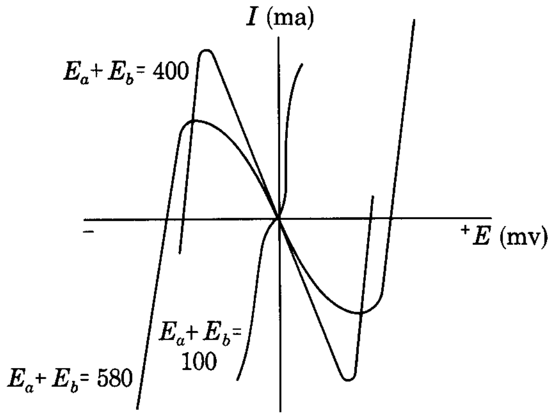
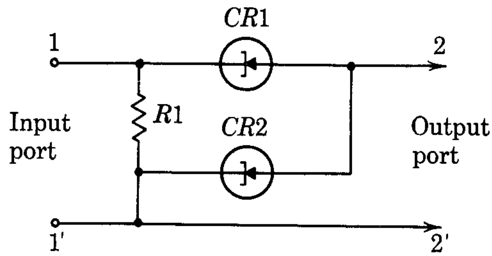


FIG. 6–15. Family of curves of coupled pair

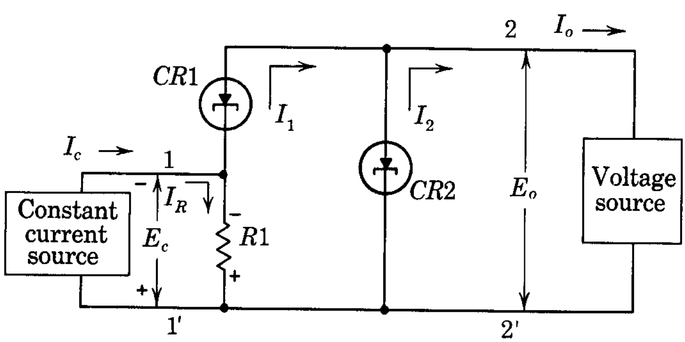
6–8. Controlled Negative Resistance, Two Tunnel Diodes

A controlled-negative-resistance *device* can be obtained by connecting two tunnel diodes and a resistor, as shown in Fig. 6–16A. The arrangement can be considered a single two-port device; the negative resistance displayed at the output port (terminals 2–2′) depends on the *control* current introduced at the input port (terminals 1–1′). As in the case of the coupled pair (par. 6–7), a *family* of characteristic curves can be obtained. Thus a fixed control current can be introduced at the input port while the output port voltage is varied from zero to its desired maximum value. The output port voltage and current are measured and plotted to form one characteristic curve of the *family.* The control current is then fixed at another value and the procedure repeated to produce another characteristic curve of the *family.* The procedure is repeated as necessary. A *typical* family of curves is shown subsequently in Fig. 6–20. An analysis of how the curves are generated is given in *a* below. Discussion of possible applications of such a device is given in *b* below.

a. Figure 6–16B shows the same circuit rearranged. In addition a control current (I_c) is introduced at the input port from a *constant-current source.* Control current I_c divides into two portions; current I_R through resistor $R1$ and current I_1 through diode $CR1$. At an output port an ideal (no internal resistance) voltage source is connected. The output port voltage (E_o) is directly across diode $CR2$ and therefore determines directly the current flow



A. Typical arrangement



B. Biased circuit

FIG. 6–16. Controlled-negative-resistance *device* using two tunnel diodes and a resistor

(I_2) through diode $CR2$. The output current (I_o) is the sum of current I_1 and I_2. The analysis divides itself into two parts; determination of current I_1 at different values of E_o with a fixed control current I_c, and addition of I_1 and I_2. For this analysis diodes $CR1$ and $CR2$ are assumed to be identical, each having the diode curve shown subsequently in Fig. 6–18, which indicates a peak current of 0.2 ma. Resistor $R1$ has a value of 250 ohms and the control current is fixed at 0.40 ma.

1. Initially when voltage E_o is zero, the circuit is reduced to that shown in Fig. 6–17A. Diode $CR2$ is shorted out of the circuit and resistor $R1$ and diode $CR1$ are in parallel across the constant-current source. To determine

the division of current between $R1$ and $CR1$, it is necessary to resort to the technique used in Chapter 3 to demonstrate graphically the current gain of a parallel amplifier; i.e., the load line (250 ohms) is drawn on the same chart as the diode curve (Fig. 6–18). The current in resistor $R1$ and diode $CR1$ are then added point by point to yield the composite curve (dashed line). The control current value (0.40) is located at the vertical axis and

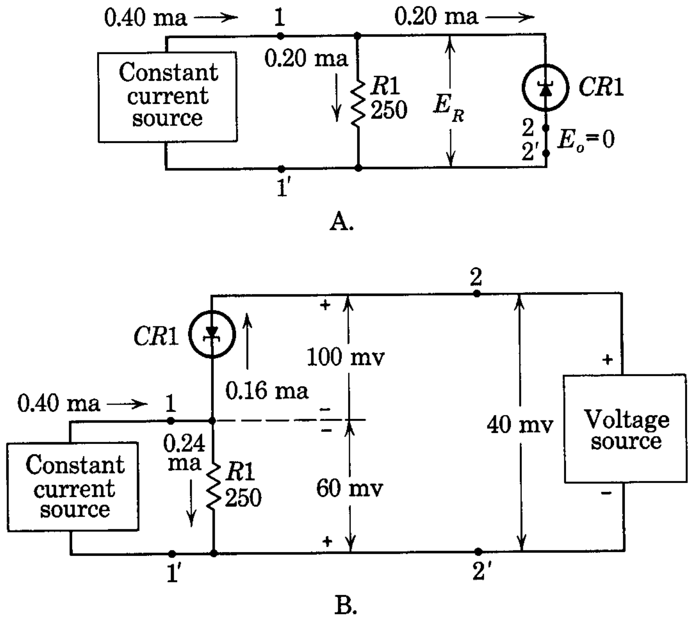


FIG. 6–17. Circuits used to determine output voltage and diode $CR1$ current relationship in controlled-negative-resistance *device*

a horizontal line drawn through it until it intersects the composite curve. At the point of intersection a vertical line is drawn until it intercepts the load line and the diode curve. In this case the point of intersection shows 0.20 ma through the diode and 0.20 ma through the resistor with 50 mv across each. (Note that if the control current were other than 0.40 ma, the current would not divide equally between resistor and diode.) All the values determined for the initial condition are entered in a table, such as Table 6–1; the columns are diode $CR1$ voltage (E_{D1}), diode $CR1$ current (I_1), resistor $R1$ current (I_R), resistor $R1$ voltage (E_R), and output port voltage (E_o). The purpose of the table is to obtain a list of values for I_1 and E_o so that a curve can be drawn for these two values and added to the curve for I_2 and E_o. To obtain additional values of I_1 and E_o, consider the circuit shown in Fig. 6–17B. Assume a voltage of 100 mv across diode $CR1$; by

using the diode curve (Fig. 6–18), it can be seen that the current through the diode is 0.16 ma. The current through the resistor must be the control current (0.40 ma) less the diode current (0.16 ma) or 0.24 ma. By Ohm's

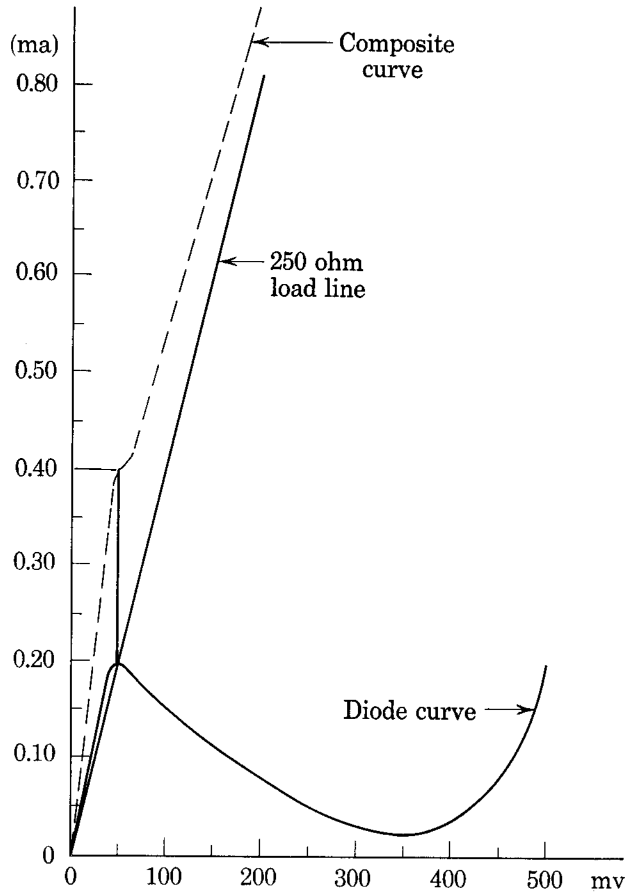


FIG. 6–18. Graphical method used to determine initial division of control current in controlled-negative-resistance *device*

law ($I \times R$) the voltage across the resistor is calculated to be 60 mv. Because resistor $R1$ voltage and output voltage (E_o) are series aiding (must add to E_{D1}), E_o voltage is obtained by subtracting E_R from E_{D1} and is found to be 60 mv. These values are also entered in the table. The process is repeated for the indicated values (arbitrarily chosen) of E_{D1}.

2. The values determined for E_o and I_1 are plotted as indicated (Fig. 6–19). Note that the peak and valley currents of diode $CR1$ occur 255 mv

TABLE 6–1

E_{D1} (mv)	I_1 (ma)	I_R (ma)	E_R (mv)	E_o (mv)
50	0.20	0.20	50	0
100	0.16	0.24	60	40
150	0.12	0.28	70	80
200	0.08	0.32	80	120
350	0.02	0.38	95	255
400	0.04	0.36	90	310
450	0.08	0.32	80	370
500	0.02	0.20	50	450

apart with respect to E_o, instead of 300 mv (350 − 50) for *CR*2. This is caused by the control current. In other words, the control current does not only voltage-offset the diode *CR*1 curve with respect to E_o but also varies its shape. Current I_o (dashed line) is obtained by adding the I_1 curve and the I_2 curve. Note that the valley current of I_o is approximately 0.07 ma

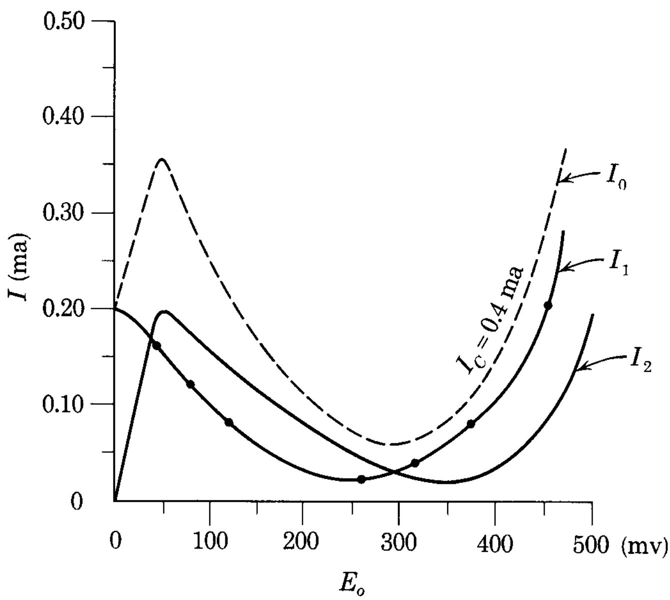


FIG. 6–19. Controlled-negative-resistance device output current-voltage curve with fixed value of control current

and occurs at 300 mv. If the same procedure is followed for a control current of 0.60 ma instead of 0.40 ma, it will be found that the composite valley current is raised to 0.08 ma and occurs at 250 mv. In other words, the greater the control current, the higher the total valley current, and the smaller the peak-to-valley current ratio. Since the negative resistance is inversely proportional to this ratio, the negative resistance is increased with

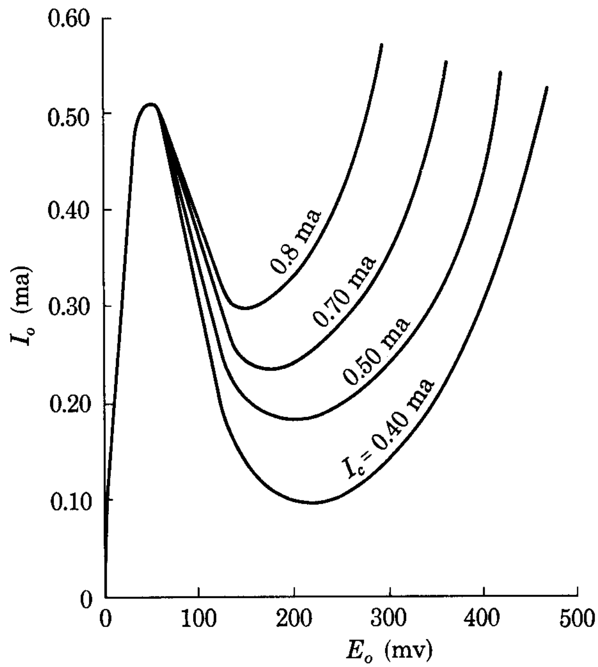


FIG. 6–20. Controlled-negative-resistance *device* (using two tunnel diodes), characteristic curves

increased control current and decreased with decreased control current. The net result is a controlled-negative-resistance device. The family of curves in Fig. 6–20 are curves obtained under practical conditions for two diodes each having a peak current of 0.25 ma and separated by a resistor of approximately 200 ohms. Under practical conditions the current source and the voltage source are not *ideal* devices, so that direct graphical derivation of these curves would be complex.

3. Characteristic curves of various shapes and possible applications can be obtained by using different values of resistance, as well as diodes made of different materials or made of the same material but having different peak currents, or peak-to-valley current ratios.

b. Possible uses of the controlled-negative-resistance device are:

1. Variable gain amplifier. The negative resistance displayed at the output port can be utilized for amplification, while the gain is varied manually or by AVC (automatic volume control) by increasing or decreasing the control current introduced at the input port. Increased control current would decrease the gain (minimum peak-to-valley current), the decreased control current would increase the gain (maximum peak-to-valley current).
2. Amplitude-modulated oscillator. The negative resistance at the output port can be used for high-frequency oscillation, whereas the input control current is varied at an audio rate. The resultant output is an amplitude-modulated signal. Such a circuit is covered in Chapter 9.
3. Variable swing relaxation oscillator. A relaxation oscillator, such as a multivibrator, can be built by using the negative resistance at the output port. The output waveform can be varied in amplitude in accordance with a variable input at the input port; the input waveform can be a sawtooth, a sinewave, a square wave, or any other form.

6–9. Controlled Negative Resistance, Three Tunnel Diodes

Three tunnel diodes and two resistors can be connected as shown in Fig. 6–21, and, as in the case of the two tunnel diode connection (par. 6–8), the arrangement can be considered a single two-port device. The negative resistance displayed at the output port (terminals 2–2′) depends on the *control* current introduced at the input port (terminals 1–1′). The analysis of the three tunnel diode arrangement is similar to the analysis for the two tunnel diode arrangement. A *family* of characteristic curves for the three tunnel diode arrangement is shown in Fig. 6–22. Three identical tunnel diodes were used, each having a peak current of 0.37 ma; resistors $R1$ and $R2$ are equal-value resistors each having a value of approximately 300 ohms.

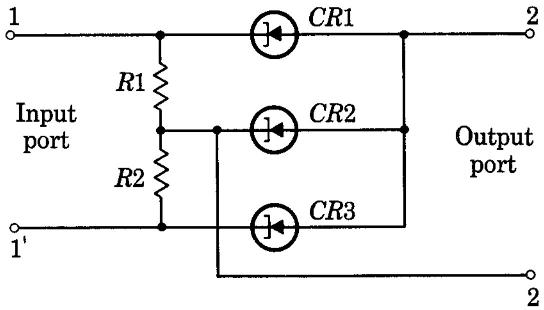


Fig. 6–21. Controlled-negative-resistance *device* using three tunnel diodes and two resistors

The three tunnel diode arrangement permits greater variation in minimum to maximum negative resistance by the control current; also a greater power-handling capacity is indicated. All types of applications possible for the two tunnel diode arrangement (par. 6–8*b*) are also possible for the three tunnel diode arrangement. The main disadvantage of the three tunnel diode arrangement is that no common ground can be established for the

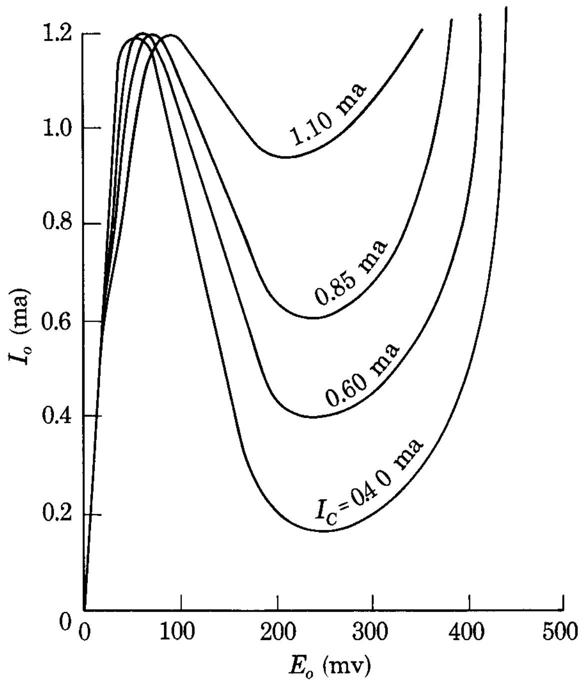


FIG. 6–22. Characteristic curves of three tunnel diode, controlled-negative-resistance *device*

input and the output circuit; one input and one output lead are at the same potential in the two tunnel diode arrangement.

6–10. Constant-Current Source, Limited Voltage Range

A good approximation to a constant-current source can be obtained by placing a tunnel diode ($CR1$) in parallel with a resistor (R_p) (Fig. 6–23A). The source voltage (E_B) must be a fixed value. For the moment consider load resistor R_L shorted. Figure 6–23B plots the current through the parallel resistor (R_p) and the tunnel diode ($CR1$). The composite curve (I_T) is the sum of the two currents. Note that if resistor R_p value is properly chosen

for the particular tunnel diode, the total current indicated between points 1 and 2 represents a relatively fixed value; i.e., for this portion of the composite curve the circuit arrangement acts like a constant-current source. Load resistor (R_L), placed in the circuit as shown (Fig. 6–23A), will experience a constant current provided the load-resistance values used limit the voltage applied (E_A) to resistor R_p and diode $CR1$ within points 1 and 2 on the composite curve. A constant-current source employing tunnel diodes and having a more extended voltage range is discussed in paragraph 6–12.

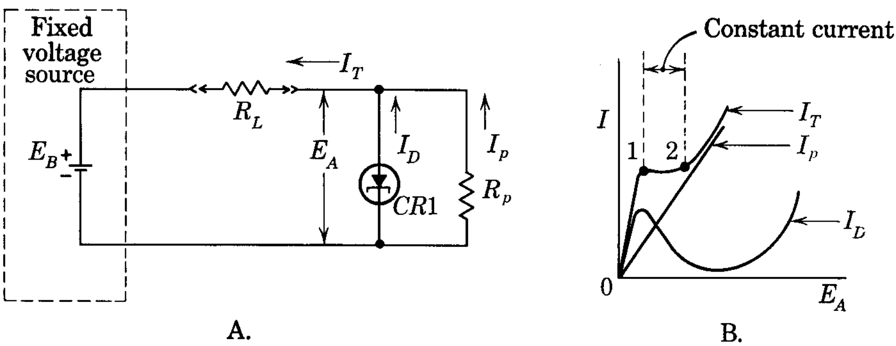


Fig. 6–23. Tunnel diode circuit (and current-voltage characteristic) used to produce a limited-voltage, constant-current source

6–11. Negative Valley-Current Characteristic

a. Figure 6–24A shows an arrangement of two tunnel diodes and two batteries which results in a composite characteristic having a negative valley current. Diode $CR1$ is connected directly across battery E_A; the only voltage measurable across diode $CR1$ is battery voltage E_A. Diode $CR2$ bias is the sum of voltages E_A and E_V. Both diodes are made of the same semiconductor material and therefore have the same peak and valley voltages. Battery voltage E_V is equal to the valley voltage and is considered a fixed voltage. The composite characteristic and the individual characteristic of each diode is obtained by varying voltage E_A. The individual characteristics are positioned so that their instantaneous currents can be added to produce the composite current.

b. Figure 6–24B shows the current-voltage characteristic of diode $CR1$ across which voltage E_A is directly applied. Note that voltages E_A and E_V are opposing voltages. When voltage E_A equals voltage E_V, there is *zero* voltage across diode $CR2$ (Fig. 6–24C), whereas diode $CR1$ is forward biased at its valley point. An increase in voltage E_A *reverse* biases diode $CR2$ and forward biases diode $CR1$ into the injection region. When volt-

age E_A equals zero, diode $CR1$ current is zero, but diode $CR2$ is forward biased at its valley point. Reversal of the polarity of voltage E_A reverse biases diode $CR1$, while diode $CR2$ is forward biased into its injection region. Because the diodes are placed in the circuit with the cathode of one connected at the same point as the anode of the other, the characteristic of diode $CR2$ must be inverted with respect to the characteristic of diode $CR1$ as shown.

c. At this point the composite characteristic (Fig. 6–24D) can be obtained by adding the currents in diodes $CR1$ and $CR2$ to obtain the total current (I_T) for each value of voltage E_A. Note that the composite characteristic displays a *negative* valley current while the applied voltage (E_A) polarity is in one direction only. One application of this arrangement produces a constant-current source having an extended voltage range (par. 6–12).

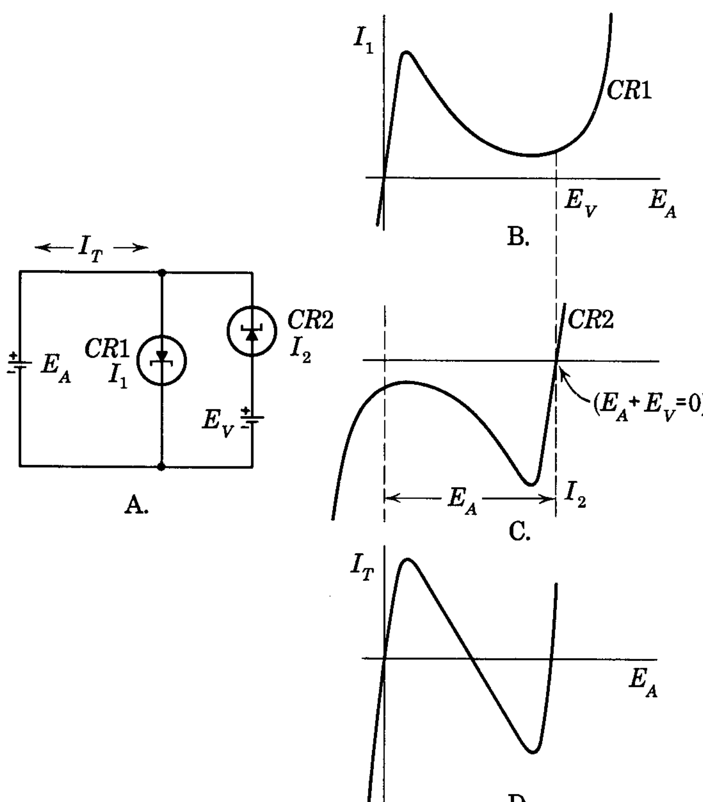


FIG. 6–24. Parallel arrangement of tunnel diodes to produce a negative valley-current characteristic

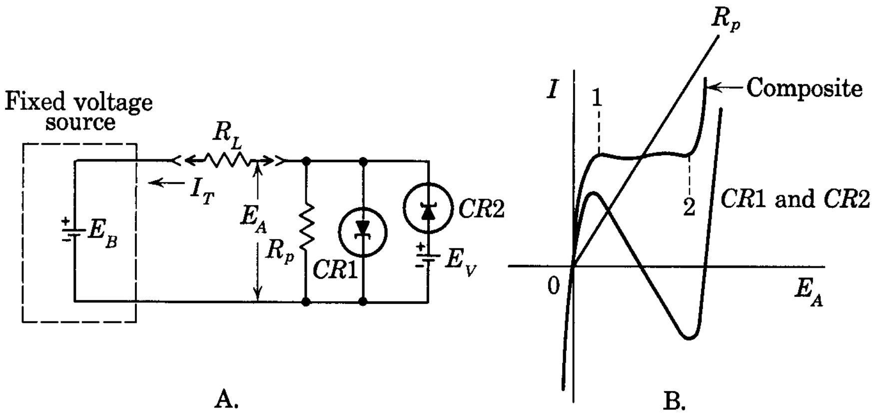


FIG. 6–25. Tunnel diode circuit (and current-voltage characteristic) used to produce an extended voltage, constant-current source

6–12. Constant-Current Source, Extended Voltage Range

A constant-current source having a limited voltage (and load resistor) range was discussed in paragraph 6–10. A circuit arrangement which produces a constant-current source having a larger voltage range is shown in Fig. 6–25A. Diodes $CR1$ and $CR2$ and battery E_V produce the negative valley-current characteristic discussed in paragraph 6–11 and shown in Fig. 6–25B. The current drawn by parallel resistor R_p is shown as a straight line. The total current (I_T) drawn from battery E_B is the sum of the currents through parallel resistor R_p and the diode pair. Current I_T is shown as the composite curve. If resistor R_p is properly chosen, the portion of the composite curve between points 1 and 2 will represent a relatively constant current. A load resistor (R_L) can be placed in the circuit as shown so that current I_T flows through it. With a fixed voltage source (E_B) the load-resistor value can vary over a wide range while the current through it remains constant. The load resistor can vary within a range that limits the voltage applied (E_A) to the parallel combination of resistor R_p and diodes $CR1$ and $CR2$ within points 1 and 2 on the composite curve.

SECTION III. BACKWARD AND BACK DIODES

6–13. General Properties

The back and backward diodes are the result of research and development work on the tunnel diode. The tunnel diode junction is formed by two heavily doped regions of p-type and n-type semiconductor materials. This combination results in a very narrow junction barrier across which

current carriers can tunnel (Chapter 2). Figure 6–26A shows the resultant characteristic of such a junction. If the percentage of doping in one or both semiconductor regions is reduced, the peak current is reduced (Fig. 6–26B). Further reduction of doping results in a characteristic (Fig. 6–26C) in which the current in the forward direction (at peak) is less than the current in the reverse direction for the same magnitude of applied voltage. Continued reduction in percentage of doping results in complete elimination of the peak current (Fig. 6–26D).

a. Back Diode. A diode having a characteristic such as that shown in Fig. 6–26C is referred to in this text as a *back* diode. It is a diode having a very low peak current but a negative resistance that is still usable in certain circuits for amplification or oscillation. Actually the negative resistance of the back diode is much higher than that of the tunnel diode. This fact is readily observed by studying Fig. 6–27. Figure 6–27A shows the tunnel diode curve with ΔI and ΔE indicated; Fig. 6–27B shows the back diode curve similarly marked. The negative resistance in each case is the ratio of ΔE to ΔI. ΔE is the same value for each if each is made of

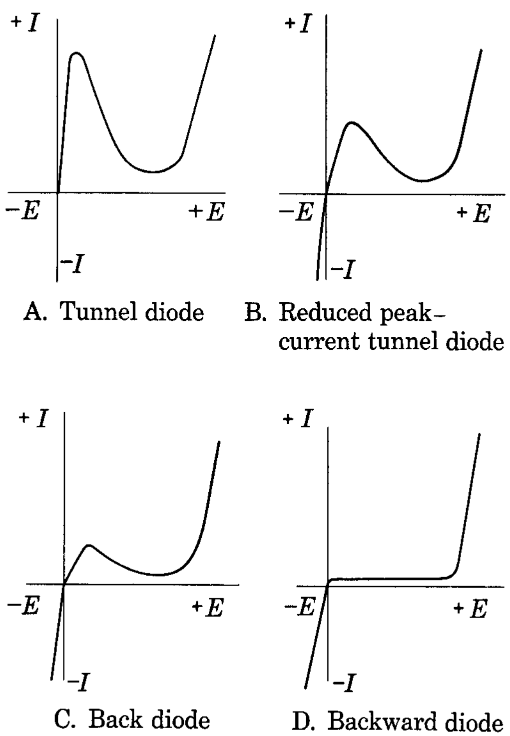


FIG. 6–26. Characteristic curves showing transition from tunnel diodes to back and backward diodes

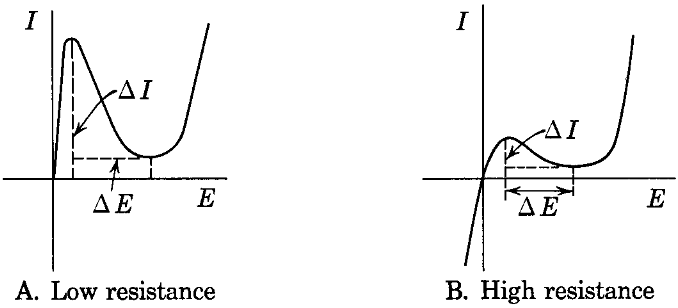


FIG. 6–27. Comparison of negative resistance of tunnel and back diodes

the same semiconductor material. ΔI is much smaller for the back diode; therefore its negative resistance is very high. Back diodes have a negative resistance of 600 to 4000 ohms, whereas tunnel diodes have a negative resistance of 40 to 150 ohms. Because of the low current, the power-handling capacity of the back diode is much less than that of the tunnel diode. Oscillator circuits employing the back diode are discussed in Chapter 7.

b. Backward Diode. A diode having a characteristic such as that shown in Fig. 6–26D is referred to in this text as a *backward* diode. It is a diode having a zero or negligible peak current. Much of the technical literature uses the terms back and backward diodes interchangeably, which can cause confusion. Use of the backward diode to modify the characteristics of the tunnel diode is discussed in paragraphs 6–14 and 6–15.

c. Graphical Representation. The graphical representation of the back or backward diode is shown in Fig. 6–28. It is the symbol of a rectifying diode with an angle mark (L) within the circle. The anode is made of p-material and the cathode is made of n-material.

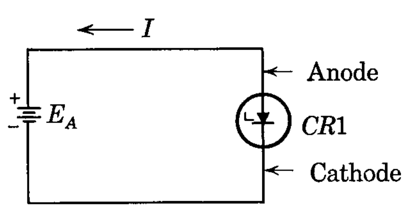
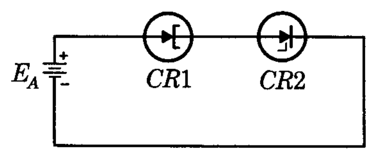


FIG. 6–28. Graphical representation of back and backward diodes

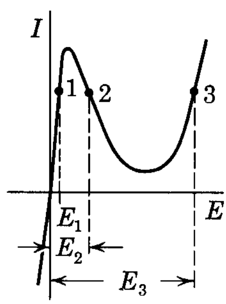
6–14. Negative Resistance at Higher Voltage

a. The tunnel diode and the backward diode can be placed in series as shown in Fig. 6–29A so that the normal negative-resistance slope of the tunnel diode occurs at a higher voltage. Note that both diodes are biased in the forward direction. Figure 6–29B shows the characteristic of the tunnel diode alone with three points of equal current flow arbitrarily chosen and marked 1, 2, and 3. The required voltage drops across the tunnel diode are marked E_1, E_2, and E_3,

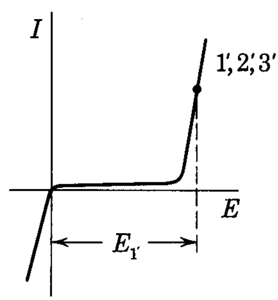
respectively. The same current flows through the backward diode, but for the three points on the tunnel diode curve only one point (1′, 2′, and 3′) occurs on the backward diode curve (Fig. 6–29C). The voltage drop across the backward diode is the same for the three conditions and is marked E_1'. The composite characteristic is shown in Fig. 6–29D. The same three current points are marked 1, 2, 3. The total voltages applied are equal to the respective voltage drops across the tunnel diode offset (increased) by the voltage drop across the backward diode. The overall effect is to voltage-shift the negative-resistance region of the tunnel diode by the voltage drop required by the backward diode.



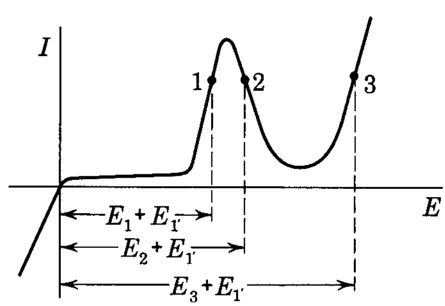
A. Tunnel and backward diodes in series



B. Tunnel diode curve

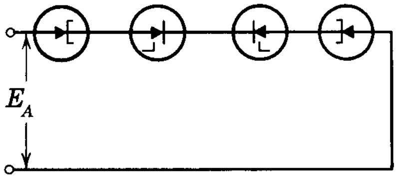


C. Backward diode curve

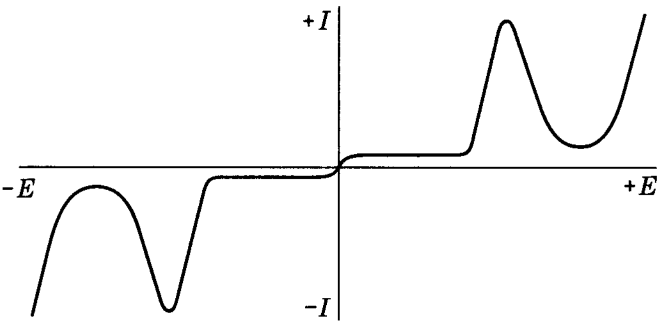


D. Composite curve

FIG. 6–29. Tunnel diode and backward diode showing individual and composite characteristic curves



A. Circuit arrangement



B. Composite curve

FIG. 6–30. Tunnel diodes and backward diodes arranged to produce a higher-voltage symmetrical composite characteristic

b. Figure 6–30A shows one pair of (tunnel and backward) diodes that would be forward biased for a particular value of applied voltage in series with another pair that would be reverse biased for the same voltage. In each case the negative-resistance region of the particular tunnel diode would be voltage-shifted by the voltage drop across the backward diode forward biased simultaneously. The composite characteristics of the four diodes are shown in Fig. 6–30B. Diodes *CR*1 and *CR*2 account for that portion of the curve to the right of the vertical axis; diodes *CR*3 and *CR*4 account for that portion of the curve to the left of the vertical axis. Such a composite curve can be used in a symmetrical switching circuit requiring higher voltages in the injection bias regions.

SECTION IV. TUNNEL DIODES AND TRANSISTORS

6–15. Increased Tunnel Diode Operating Current Level

a. The tunnel diode can be used in conjunction with the transistor to increase many times the operating current of the diode. A circuit arrange-

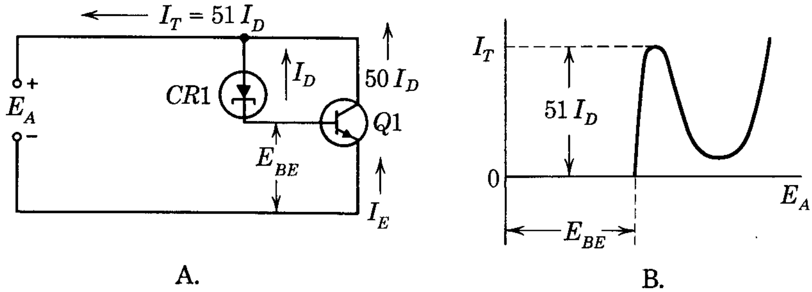


FIG. 6–31. Tunnel diode and transistor (with composite characteristic) arranged to increase tunnel diode current

ment designed to achieve this effect is shown in Fig. 6–31A. Diode $CR1$ is in series with the base-emitter junction of *n-p-n* transistor $Q1$ and determines the transistor base-current characteristic. Transistor $Q1$ is used in the common-emitter configuration. The current amplification factor of the common-emitter configuration varies between 30 and 50 times, depending on the particular transistor used. In this case it is assumed that the current amplification factor of $Q1$ is 50; i.e., the collector current is 50 times the base current. The electron current flow through each lead is indicated by the arrows. Note that the collector current is 50 times the diode and base current (I_D). The collector current and diode current add to the common connection so that the total current is 51 times the diode current alone.

b. Figure 6–31B shows the composite current-voltage characteristic. Note that no current flows until the base-emitter voltage (E_{BE}) is sufficiently large to turn on the transistor. This condition offsets the normal tunnel diode voltage by the magnitude E_{BE}. The condition can be overcome by placing a separate battery voltage in the base lead (Fig. 6–32A) or the emitter lead (Fig.

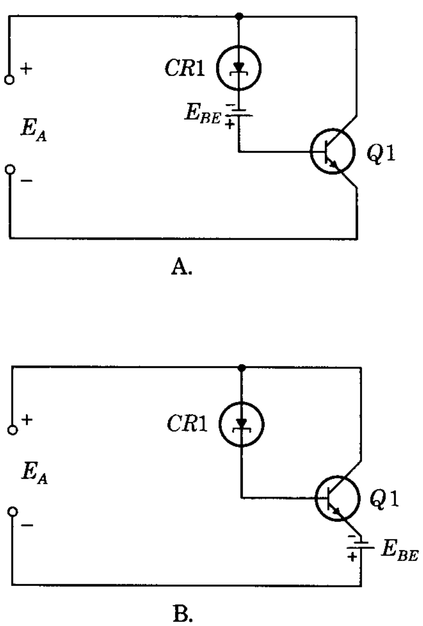


FIG. 6–32. Battery in base lead or emitter lead used to overcome base-emitter offset voltage

6–32B). The separate battery voltage (E_{BE}) places the transistor on the verge of conduction when the applied voltage (E_A) equals zero.

c. The circuit will operate equally well with a *p-n-p* transistor provided the battery polarities are reversed and the cathode and anode of the tunnel diode are reversed in position.

d. The cutoff frequency of the composite circuit will be limited by the cutoff frequency of the much slower acting transistor. The *average* transistor operates up to 100 megacycles, whereas the tunnel diode can operate at several kilomegacycles.

6–16. Decreased Tunnel Diode Operating Current Level

a. The tunnel diode can be used in conjunction with the transistor, with the result that the input current to the combination is many times smaller than the current drawn by the tunnel diode alone. A circuit arrangement designed to achieve this effect is shown in Fig. 6–33A. Diode *CR*1 is in series with the emitter lead of *n-p-n* transistor *Q*1. The emitter current and the diode current (I_D) are the same. In the average transistor about 98% of the emitter current ($0.98I_D$) reaches the collector and 2% of the emitter current ($0.02I_D$) reaches the base. In this example, then, the base current is 50 times less than the tunnel diode current. Such a circuit can be used to drive the diode from a high-resistance, very low current source.

b. A plot of the base current versus the applied voltage is shown in Fig. 6–33B. Note that no current flows until the base-emitter junction voltage (E_{BE}) is large enough to turn on the transistor. This offset voltage can be overcome by inserting a battery voltage (E_{BE}) in the emitter lead at the terminals marked 1 and 2. The negative terminal of the battery must be placed at point 1. This battery voltage places the transistor on the verge of conduction when applied voltage E_A equals zero.

c. Note also that the output voltage can be taken from the emitter or the collector. The emitter voltage is exactly the diode voltage and is very low. Depending on the particular transistor and the collector load resistor (R_L), the collector voltage can be many times the diode voltage.

d. A *p-n-p* transistor can be used in the circuit provided the battery voltage is reversed and the positions of the diode cathode and anode are interchanged.

6–17. *S*-type and *N*-type Characteristic Curves

Devices that display negative resistance, decreasing current with increasing voltage, are referred to as either *S*-type or *N*-type negative-resistance devices. These designations refer to the general shape of their voltage-current characteristic curve when voltage is plotted vertically and current horizontally.

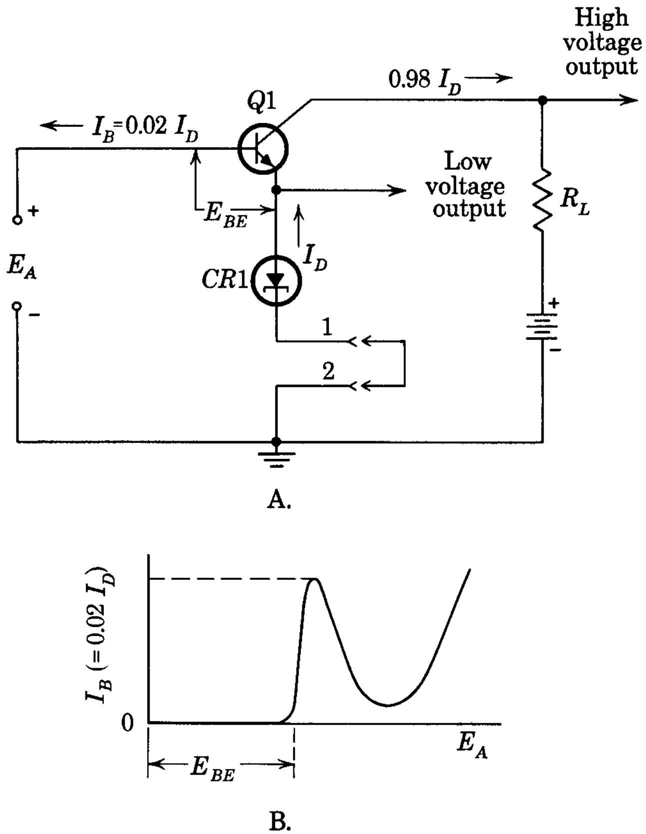


FIG. 6–33. Circuit arrangement (and composite characteristic) used to decrease the tunnel diode input current

a. An idealized *S*-type curve is shown in Fig. 6–34A. Devices that display this curve are the tunnel diode, the tetrode vacuum tube (used as a transitron oscillator, par. 3–1), and the point-contact transistor used in the common-emitter configuration. Note that within the negative-resistance range, for any given current (dashed line), three possible values of voltage are possible. For any given value of voltage, however, only one current value is possible. The *S*-type characteristic device, therefore, is said to be voltage stable, or voltage controlled; i.e., its operating point is determined by a stiff (constant) voltage source when operated as an oscillator or amplifier.

b. An idealized *N*-type curve is shown in Fig. 6–34B. Devices that display this curve are the common-base configuration of the point-contact transistor (par. 3–3), the unijunction transistor (par. 3–5), the four-layer

diode (par. 3–4), and gaseous discharge tubes. Note that within the negative-resistance range for any given voltage (dashed line), three values of current are possible. For any given value of current, however, only one voltage value is possible. The N-type characteristic device, therefore, is said to be current stable, or current controlled; i.e., its operating point is determined by a constant-current source when operated as an oscillator or amplifier.

c. The current-voltage characteristic of the tunnel diode which is normally an S-type curve can be transformed into an N-type curve when the tunnel diode is combined in a particular arrangement with a transistor (par. 6–18).

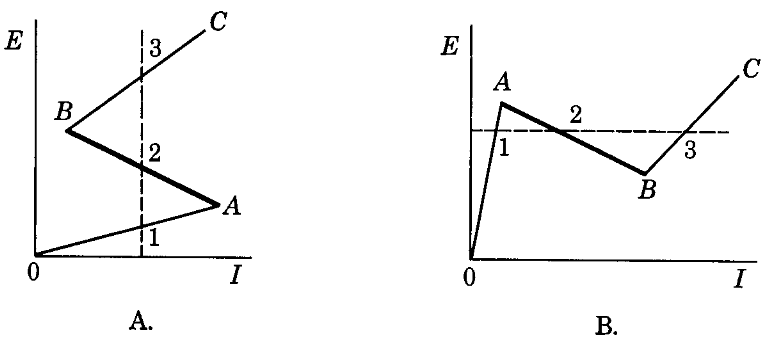


FIG. 6–34. Idealized S-type and N-type negative-resistance characteristic curves

6–18. Tunnel Diode and N-type Characteristic

a. The tunnel diode can be combined with a transistor (Fig. 6–35A) to produce an N-type voltage-current characteristic curve (par. 6–17). Tunnel diode $CR1$ is placed in parallel with the base emitter of transistor $Q1$. Current limiting resistor $R1$ is placed in series with the parallel arrangement. The bias source consists of battery E_B in series with a heavy resistor R_s. To develop the composite characteristic curve (Fig. 6–35B), assume switch $S1$ is closed.

1. Voltage and current start to rise from zero to a steady-state condition. From point 0 to point A the transistor does not conduct because its base-emitter voltage is too low. The only current flow is through resistor $R1$ and diode $CR1$. The applied voltage (E_A) equals the peak voltage of the tunnel diode plus the peak current of the diode multiplied by the resistance of $R1$.

2. At point A the diode switches rapidly to its high-voltage state and forward biases transistor $Q1$ sufficiently so that it conducts. The increase

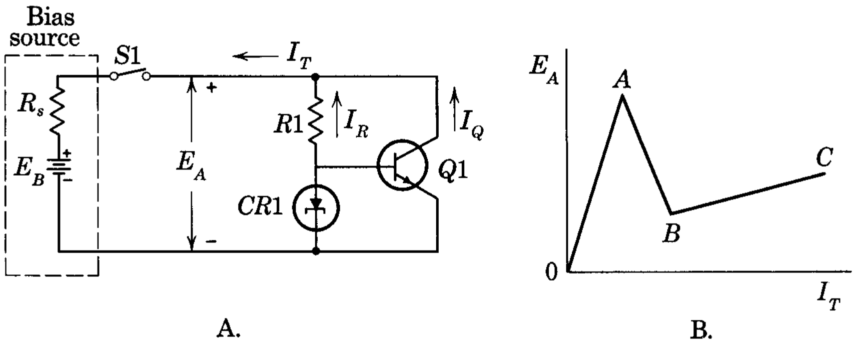


FIG. 6–35. Tunnel diode and transistor arranged to produce an N-type composite characteristic curve

in total current (I_T) due to the contribution of collector current by $Q1$ drops the applied voltage (E_A) to the low value at point B because of the increased voltage drop across resistor R_s. The current (I_T) increases to point C without appreciable increase in voltage E_A because the emitter collector resistance of transistor $Q1$ decreases with increasing current.

b. For this arrangement to be used as a switching device, it is most desirable that the voltage at point B be as low as possible. The voltage at point B is determined mainly by the tunnel diode valley-current flow through resistor $R1$. If a separate source is provided to supply the valley current to diode $CR1$, the voltage at point B can be substantially reduced. Such a circuit arrangement is shown in Fig. 6–36A. Battery E_{BB} and current limiting resistor $R2$ have been added to supply only the valley current to diode $CR1$. The composite characteristic curve is shown in Fig. 6–36B. Note the very low value of voltage E_A at point B on the curve. A relaxation oscillator employing this arrangement is covered in Chapter 7.

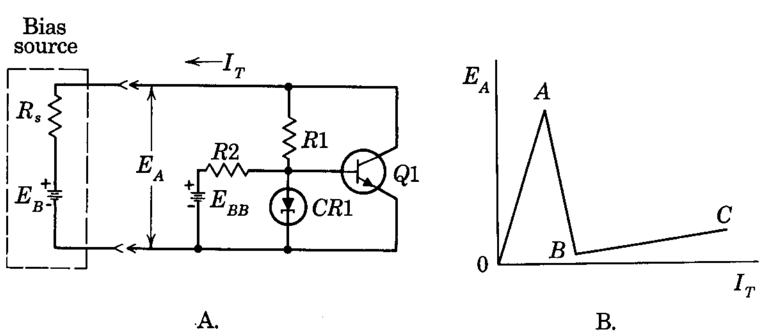


FIG. 6–36. Separate valley-current source provided to reduce the applied voltage with transistor conducting

c. The circuit discussed employs an *n-p-n* transistor. A *p-n-p* transistor can be used provided all voltage polarities and the diode are reversed.

6–19. Summary

a. Two tunnel diodes can be connected back-to-back (Fig. 6–3) to produce a negative conducting region for positive or negative applied voltages.

b. Any number of tunnel diodes can be placed in series (Fig. 6–4) to produce a composite characteristic useful in applications requiring multistate switches.

c. The tunnel diode, in parallel with a load resistor and fed from a constant-current source (Fig. 6–8), can approximate a relatively constant-voltage source.

d. The tunnel diode coupled pair (Fig. 6–9) results in a two-port *device* with a current-voltage characteristic for one port that can be varied by changing the voltage at the other port (Fig. 6–15). The current-voltage characteristic will display negative resistance if the total bias voltage is greater than twice the diode peak voltage and less than twice the diode valley voltage.

e. The resistances of the two positive slope regions of the coupled-pair characteristic are equal.

f. A controlled-negative-resistance *device* can be obtained by using a resistor and two tunnel diodes (Fig. 6–16), or by using two resistors and three tunnel diodes (Fig. 6–21). Either arrangement can be used in circuits such as variable gain amplifiers, variable swing relaxation oscillators, and amplitude-modulated oscillators. The main advantage of the two-diode over the three-diode arrangement is that the former permits a common ground connection between the input and the output circuits.

g. The tunnel diode, placed in parallel with a suitable resistor and fed from a fixed voltage source (Fig. 6–23), can approximate a constant-current source over a limited load-voltage range.

h. Two tunnel diodes properly biased and arranged (Fig. 6–24) can produce a negative valley-current characteristic.

i. The same combination that produces the negative valley current (*h* above) when paralleled with the proper resistor (Fig. 6–25), can approximate a constant current source over an extended load-voltage range.

j. By progressive reduction in the percentage of doping materials used to make tunnel diodes, a back diode and a backward diode can be produced (Fig. 6–26).

k. The back diode has a usable but greatly reduced peak current and its negative resistance is very high (about 4000 ohms).

l. The backward diode has a negligible or zero peak current.

m. A tunnel diode and a backward diode can be used in series (Fig. 6–29)

to produce a negative resistance at a higher voltage range. Two pairs back-to-back (Fig. 6–30) produce a symmetrical composite characteristic with negative-resistance regions at higher positive or negative voltages.

n. The tunnel diode can be used with a transistor to increase the tunnel diode current (Fig. 6–31) or to decrease the input current to the tunnel diode (Fig. 6–33).

o. A negative-resistance device having an *S*-type voltage-current characteristic (Fig. 6–34) is a voltage-controlled device. A negative-resistance device having an *N*-type voltage-current characteristic is a current-controlled device.

p. The tunnel diode can be used with a transistor (Figs. 6–35 and 6–36) to transform its *S*-type voltage-current characteristic into an *N*-type characteristic.

Chapter 7

OSCILLATORS

SECTION I. GENERAL

7–1. Sinewave Oscillators

a. A detailed discussion of the manner in which the ac negative resistance of the tunnel diode may be used in specific circuits to produce sinewave oscillations is given in Sections III and V of Chapter 3. Also discussed are the formulas for the self-resonant frequency (f_s) and the resistance cutoff frequency (f_r) of the tunnel diode (par. 3–17).

b. Practical sinewave oscillators are discussed in this chapter as follows:

1. Low-frequency oscillators (1 kc to 100 mc), which normally use lumped circuit elements, are discussed in paragraphs 7–3 through 7–6.
2. Oscillators operating above several hundred megacycles, which normally use transmission line resonators, are discussed in paragraph 7–7.
3. Oscillators operating in the microwave band normally use waveguides and cavity resonators. These circuits are discussed in paragraphs 7–8 and 7–9.

7–2. Relaxation Oscillators

a. An oscillator circuit that produces a nonsinusoidal output waveform is generally classified as a relaxation oscillator. The tunnel diode relaxation oscillator uses the inherent negative resistance of the diode in conjunction with resistance-capacitance or resistance-inductance networks to provide a switching action. The charge and discharge times of the associated networks are used to produce sawtooth, square, or pulse output waveforms. Capacitors and inductors require a definite amount of time to charge or discharge through a resistor. The measure of this time, *the time constant,* is determined by multiplying the resistance by the capacitance ($R \times C$) or by dividing the inductance by the resistance ($L \div R$).

b. Relaxation oscillators are classified as astable, monostable, or bistable. The astable oscillator is a free-running oscillator and produces a continuous

output as long as bias voltage is supplied. In addition to bias voltage, the monostable and bistable oscillators require a signal (pulse or sinewave) input to produce an ac output. The conditions which determine whether a tunnel diode (or coupled pair) relaxation oscillator is astable, monostable, or bistable are discussed in paragraphs 7–10 and 7–12. Monostable and bistable oscillators (switching circuits) are discussed in Chapter 8. Astable oscillators are discussed in Section III of this chapter as follows:

1. Single diode multivibrator (par. 7–11)
2. Coupled-pair multivibrator (par. 7–13)
3. Sawtooth generator (par. 7–14)
4. Diode-transistor multivibrator (par. 7–15)

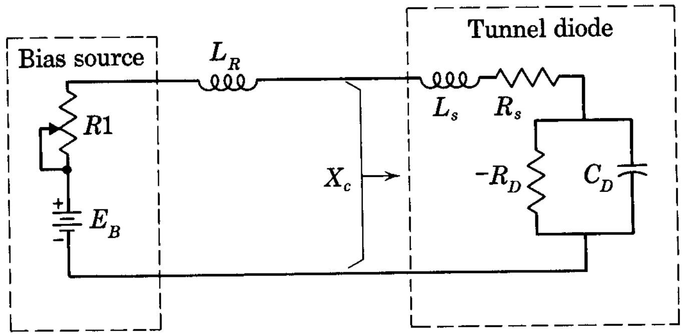
SECTION II. *LC* RESONANT OSCILLATORS

7–3. Oscillations Below and Above Self-Resonance

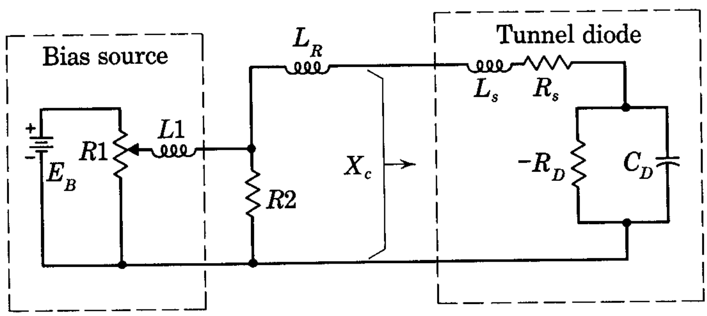
a. Below Self-resonance. 1. The equivalent circuit of the tunnel diode is shown in dashed lines in Fig. 7–1A. At the self-resonant frequency (f_s) the input impedance of the diode has zero reactance; i.e., the input impedance is resistive only. Inasmuch as at frequencies *below* self-resonance the inductive reactance of L_s decreases and the capacitive reactance of C_D increases, the input impedance displays *capacitive reactance* (X_c) as indicated. Oscillation will occur at frequencies below self-resonance if an external inductor (L_R) is added with its inductive reactance equal to the capacitive reactance (X_c). Bias is applied through variable resistor $R1$. Oscillation will occur, of course, only if the desired operating frequency is less than the resistive cutoff frequency of the diode, and the total positive resistance is less than the effective negative resistance of the diode.

2. Figure 7–1B shows the same oscillator fed from a bias source having an appreciable lead (or cable) inductance ($L1$). To decouple this inductance from the oscillator circuit, resistor $R2$ is placed in the circuit as shown. The resistance value of resistor $R2$ is ten to twenty times less than the inductive reactance of $L1$.

b. Above Self-resonance. 1. The equivalent circuit of the tunnel diode is shown in dashed lines in Fig. 7–2A. At frequencies above self-resonance, the capacitive reactance of C_D decreases and the inductive reactance of L_s increases, causing the input impedance of the diode to display *inductive reactance* (X_L) as indicated. Oscillation will occur at frequencies above self-resonance if an external capacitor (C_R) is added with its capacitive reactance equal to the inductive reactance (X_L), provided that the operating frequency is below the resistance cutoff frequency of the diode.



A. Oscillator



B. Decoupled oscillator

FIG. 7–1. Circuit arrangement for oscillator operation below self-resonant frequency of diode

2. Figure 7–2B shows the same oscillator fed from a bias source having an appreciable lead (or cable) inductance ($L1$). Resistor $R2$ is added to decouple the lead inductance in the same manner as discussed in *a*2 above.

7–4. Frequency Stability

An oscillator consisting of a tunnel diode in series with a capacitor or inductor (*a* below) has poor frequency stability compared to an oscillator using the same tunnel diode in series with a separate tank (LC) circuit (*b* below).

a. Figure 7–3 shows the series resonant circuit (Fig. 7–1A) with like elements lumped; i.e., resistor R_T represents that portion of variable resistor $R1$ that is in the circuit, plus lead and bulk resistance (R_s) of the diode, and the dc resistance (R_L) of inductor L_R; inductor L_T represents the in-

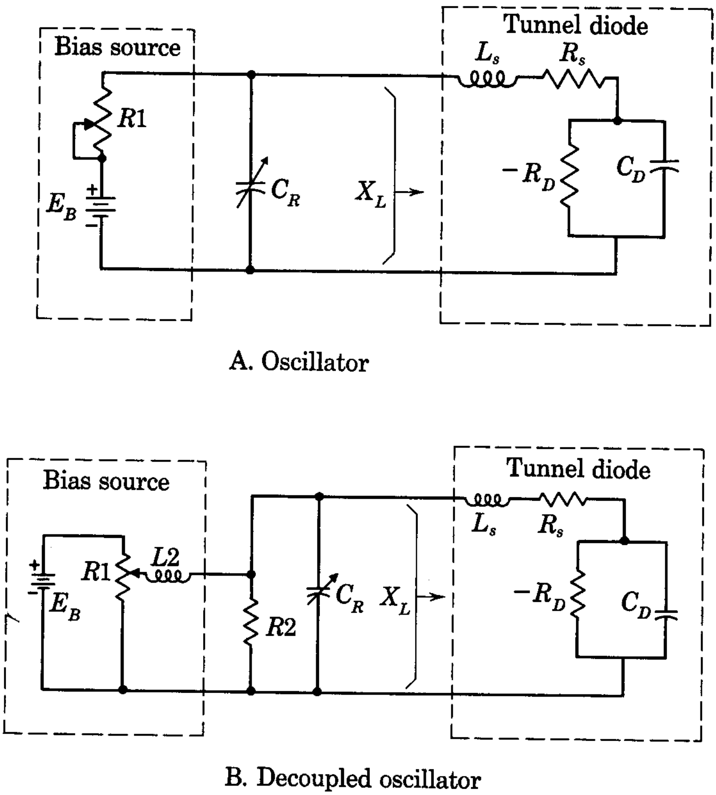
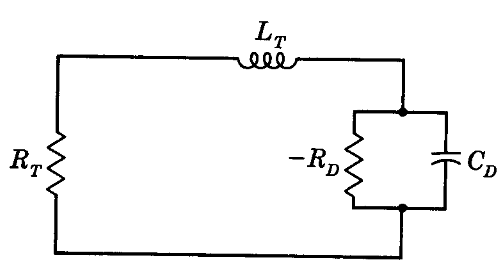


FIG. 7–2. Circuit arrangement for oscillator operation above self-resonant frequency of diode



$$R_T = R1 + R_s + R_L$$
$$L_T = L_R + L_s$$

FIG. 7–3. Circuit showing lumped constants of simple series oscillator

ductance of L_R plus the lead inductance of the diode. Resistor R_D and capacitor C_D are the remaining parameters of the diode. An analysis of this circuit results in the following formula for the oscillating frequency (f_o):

$$f_o = \frac{1}{2\pi}\sqrt{\frac{R_D - R_T}{L_T C_D R_D}}$$

This formula shows that the oscillating frequency is heavily dependent upon the negative resistance (R_D) of the diode as well as the diode capacitance (C_D). The negative resistance of the tunnel diode is not constant over the negative-resistance region of the diode (par. 4–5 and 4–6). A slight change in bias voltage results in a large change in the average value of $-R_D$ and therefore in the oscillating frequency. Variation in bias voltage also varies C_D. Temperature variations also change $-R_D$ and C_D (par. 4–7).

b. Greater frequency stability is obtained with the oscillator circuit shown in Fig. 7–4. Resistors $R1$ and $R2$ form a voltage divider to establish the bias voltage. Capacitor $C1$ parallels the tunnel diode to swamp the inherent tunnel diode junction capacitance; i.e., variations in the diode capacitance due to bias or temperature variations result in a smaller percentage change of the total capacitance across the diode. Coil $L1$ and capacitor $C2$ form a resonant tank circuit. Analysis of this circuit (ignoring diode lead resistance and inductance) results in the following formula for the oscillating frequency (f_o):

$$f_o = \frac{1}{2\pi}\sqrt{\frac{1}{L1(C1 + C2)} - \frac{1}{C1(C1 + C2)R_D{}^2}}$$

In this formula capacitance $C1$ includes the junction capacitance (C_D) of the diode. This formula shows that the oscillating frequency is heavily dependent upon the constants of the tank circuit and swamping capacitor $C1$. Practical circuits using this arrangement are discussed in paragraph 7–5.

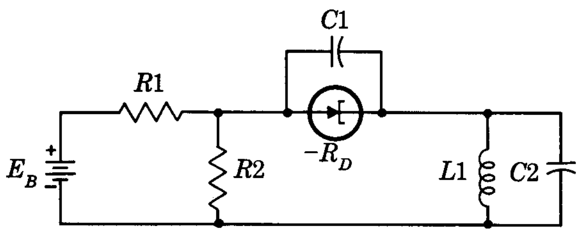
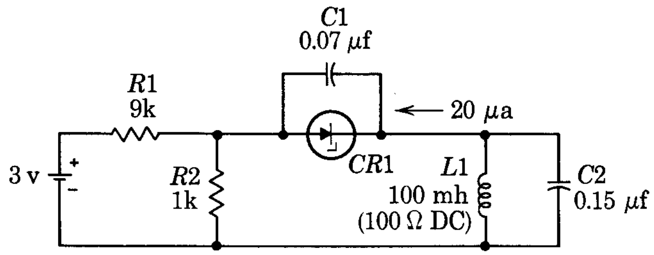


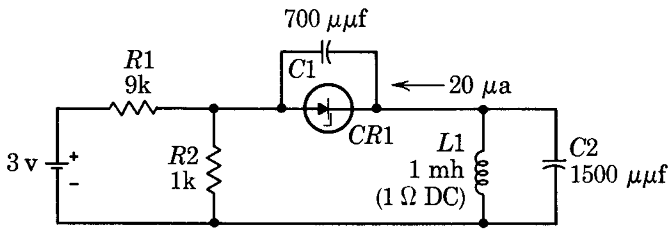
FIG. 7–4. Oscillator using tunnel diode in series with tank (LC) circuit

7-5. Series-Parallel Oscillators

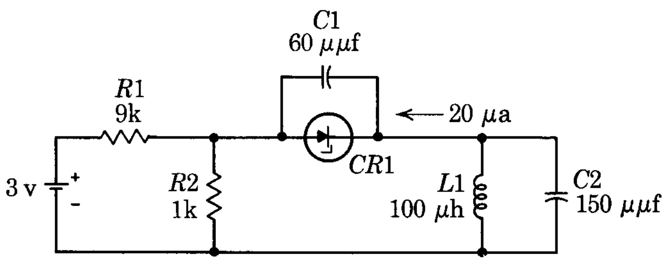
a. Back Diode Oscillators. Oscillators using a diode in series with a tank (LC) circuit are shown in A, B, and C, Fig. 7-5. The circuits oscillate



A. Back diode 1-kc oscillator



B. Back diode 100-kc oscillator



C. Back diode 1-mc oscillator

FIG. 7-5. Back diode oscillators (Adapted circuits, courtesy General Electric Company)

at 1 kc, 100 kc, and 1 mc, respectively. Capacitor $C1$ in each circuit is a swamping capacitor. The function of each part in each circuit is identical with the correspondingly referenced part of the oscillator shown in Fig. 7-4 (par. 7-4). Note that in each circuit diode $CR1$ is a back diode (par. 6-13). The negative resistance of the back diode is approximately 600 to 4000 ohms compared to 40 to 150 ohms for the tunnel diode. Such a high negative resistance permits the use of a high-resistance voltage divider ($R1$ and $R2$)

which minimizes the drain on the bias battery. Furthermore a 1-kc oscillator requires a relatively large inductance in the tank circuit. Note that the dc resistance of coil $L1$ is 100 ohms (Fig. 7–4A). The dc resistance inserted by the voltage divider equals the product of $R1$ and $R2$ divided by their sum and is less than 1000 ohms. The total positive resistance therefore is much less than the negative resistance, thereby assuring oscillation. The main disadvantage of the back diode, however, is its limited

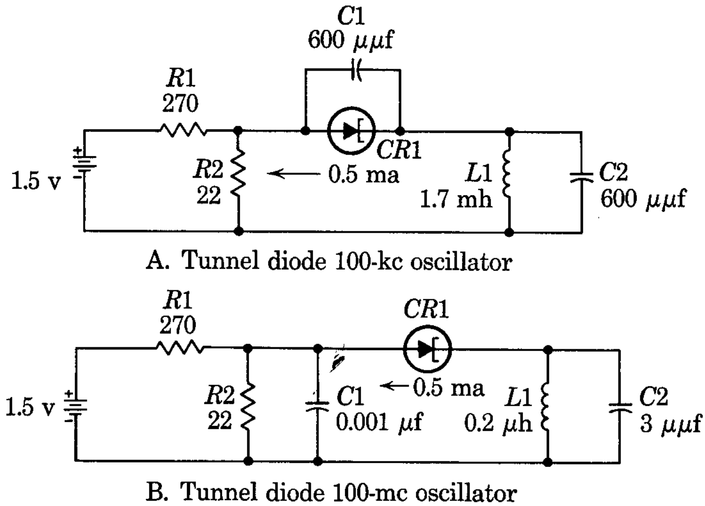


FIG. 7–6. Tunnel diode oscillators

power output; note that the current flow at bias is only 20 microamperes. For greater power output, the tunnel diode must be used (*b* below).

b. Tunnel Diode Oscillator. A 100-kc and a 100-mc tunnel diode oscillator are shown in Figs. 7–6A and B, respectively. Because of the low negative resistance of the tunnel diode, lower values of resistance are used in the voltage divider network (resistors $R1$ and $R2$). The bias current is approximately 0.5 ma. In the 100-mc oscillator no swamping capacitor parallels the diode. At 100 mc a parallel capacitor would substantially reduce the effective negative resistance of the diode (par. 3–18) so that oscillations may not occur. In addition, capacitor $C1$ ac bypasses resistor $R2$ to reduce further the positive resistance in the circuit.

7–6. Crystal Controlled Oscillator

A crystal controlled tunnel diode oscillator is discussed in *b* below. A quartz crystal (*a* below) is normally used in an oscillator circuit because

of its extremely high Q (narrow bandwidth) and good frequency stability over a given temperature range. In practical applications when the range of operating temperature or frequency tolerance is extremely critical, a constant-temperature chamber is used to house the crystal.

a. Crystal Properties. The quartz crystal graphical representation (with reference designation $Y1$), its equivalent circuit, and its impedance-frequency characteristic are shown in Figs. 7–7A, B, and C.

1. Resistance R, inductance L, and capacitance C_s in series (Fig. 7–7B) represent the electrical equivalent of the mechanical vibrating characteristic of the crystal. At series resonance of capacitance C_s and inductance L, the impedance of the crystal is low and the resonant frequency of the oscillator circuit is determined only by the mechanical vibrating characteristics of the crystal.

2. Parallel capacitance C_p represents the electrostatic capacitance between the crystal electrodes (Fig. 7–7A). Above the frequency of series resonance (1 above), the inductive reactance of inductance L is greater than

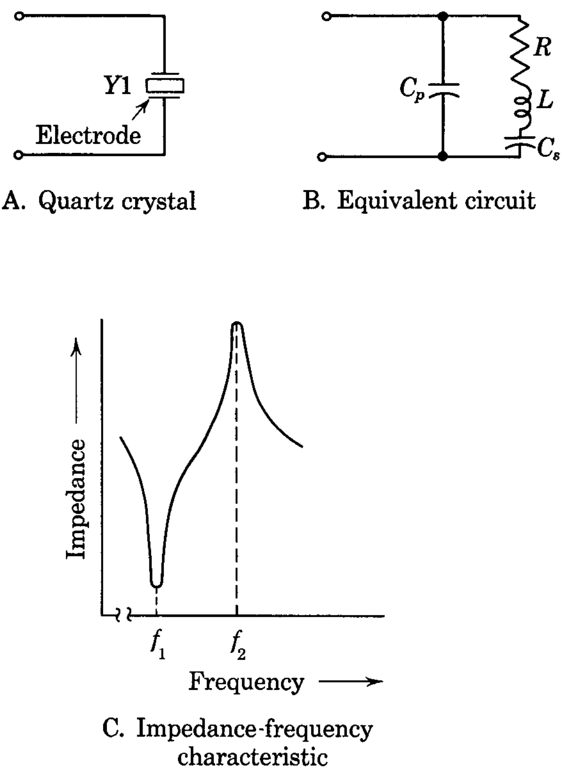
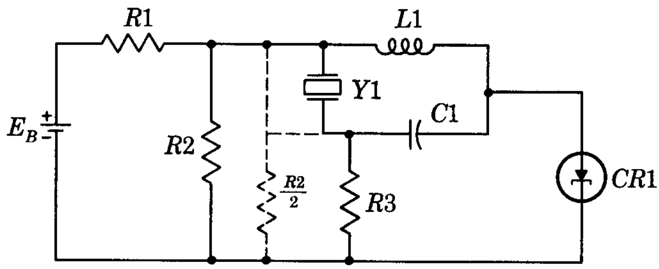


Fig. 7–7. Quartz crystal, equivalent circuit and impedance-frequency characteristic



Note: $R2 = R3$

FIG. 7–8. Crystal controlled tunnel diode oscillator

the capacitive reactance of capacitance C_s. The combination (L and C_s) appears as a net inductance. This net inductance forms a parallel resonant circuit with capacitance C_p and any circuit capacitance appearing across the crystal. At parallel resonance the impedance of the crystal is high and the resonant frequency is determined by the crystal and by externally connected circuit elements.

3. The impedance-frequency chart of a typical quartz crystal is shown in Fig. 7–7C. The lowest impedance occurs at series resonance (f_1). The highest impedance occurs at parallel resonance (f_2). The slopes of the curve are steep at each resonant point, indicating a high Q. For most crystals the difference in frequency between f_1 and f_2 is very small compared to the series resonant frequency of the crystal.

4. Either the series or parallel mode of oscillation of the crystal may be used in an oscillator circuit. The mode of oscillation is primarily determined by the impedance of the circuit to which the crystal is connected.

b. Oscillator. A crystal-controlled tunnel diode oscillator is shown in Fig. 7–8. For the present discussion ignore the dashed-line portion of the circuit.

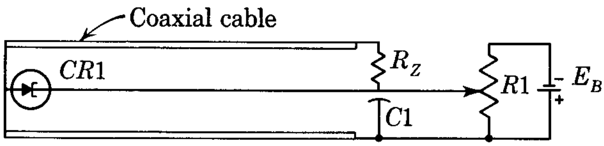
1. Bias voltage is supplied by battery E_B through voltage divider $R1$ and $R2$. Note that no dc current flows through resistor $R3$. Inductor $L1$ and capacitor $C1$ are essentially in parallel at the series resonant frequency of crystal $Y1$ when its impedance can be considered a short circuit.

2. Resistor $R2$ is selected so that its positive resistance is approximately *twice* the effective negative resistance of the tunnel diode. Resistor $R3$ is made equal to resistor $R2$. When the quartz crystal is at series resonance (low impedance), it acts as a short circuit and ac parallels resistors $R2$ and $R3$ so that the total positive resistance is one half of resistor $R2$ (dashed lines). This value equals (or is slightly less than) the diode negative re-

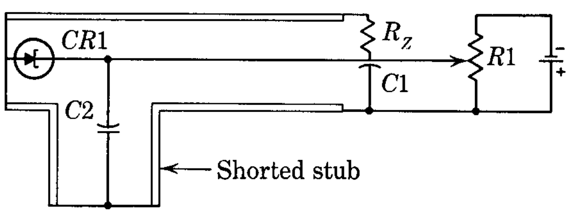
sistance and oscillations can occur. Thus the circuit can oscillate only at the series resonant frequency of crystal $Y1$.

7-7. Voltage-Tuned Oscillator

a. Figure 7-9A shows an oscillator the frequency of which can be varied from 900 mc to 1000 mc by varying the bias voltage. The circuit consists of a tunnel diode placed at the end of a coaxial transmission line and ter-



A. Voltage-tuned oscillator



B. Oscillator with shorted stub

FIG. 7-9. Tunnel diode oscillators (voltage-tuned or stabilized) using coaxial transmission lines

minated in its characteristic impedance R_Z; capacitor $C1$ is a feed-through bypass capacitor and resistor $R1$ permits variation of bias voltage (E_B). For oscillation to occur the effective negative resistance of the diode must be greater than the total positive resistance of the circuit. Because the ac impedance of the coaxial cable is relatively constant over a broad range of frequencies, the operating frequency is determined primarily by the effective negative resistance of the diode. In turn, the effective negative resistance depends upon the bias voltage applied. Figure 7-10 presents a plot of the frequency output of the voltage-tuned oscillator versus the applied diode bias voltage. The nonlinearity of this curve is caused mainly by resonances at harmonic frequencies.

b. Figure 7-9B shows the same circuit with a shorted stub cut to one quarter of the wavelength of the desired frequency (1050 mc). Capacitor $C2$ is a coaxial capacitor that prevents shorting of the dc bias. The shorted

stub is effectively a high-impedance tuned circuit in parallel with the tunnel diode. Its effect is to stabilize the operating frequency with varying bias voltage by making the frequency of operation less dependent on the effective negative resistance of the diode. The frequency output versus bias voltage for this circuit is also shown in Fig. 7–10. The output for each oscillator is obtained by inserting a loop within the coaxial cable.

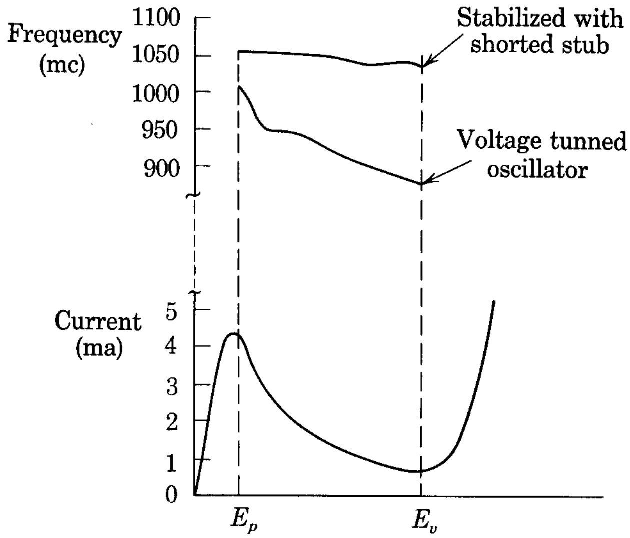


FIG. 7–10. Curves of frequency output versus bias voltage for voltage-tuned and for stabilized oscillators

7–8. Strip Transmission-Line Oscillator

a. For frequencies above 600 mc, it is common practice to use hollow waveguides as cavity resonators in amplifiers and oscillators. The physical dimensions of the hollow waveguide mainly determine the resonant frequency. For a given frequency the overall dimensions of a hollow waveguide are relatively large and can be avoided by using an alternate construction referred to as strip transmission-line construction which substantially reduces the required overall dimensions. The hollow waveguide is a completely enclosed structure except for input and output ports. The strip transmission-line construction consists of two (upper and lower) plates of conducting material separated by a low-loss dielectric; there are no side plates. The electrical field exists between the upper and lower plates; the lack of side plates leads to negligible radiation. Strip transmission-line construction is to microwave technology what printed-circuit construction is to low-frequency technology.

b. A tunnel diode oscillator using a strip transmission-line cavity resonator is shown (exploded view and assembled view) in Fig. 7–11. Within the cavity is placed a suitably constructed tunnel diode and a stabilizing resistor. Terminals are provided for introducing the dc bias and extracting the RF signal. It can be shown by mathematical analysis beyond the scope of this text that, in addition to other factors, the frequency of oscillation depends upon the positions of the tunnel diode and the stabilizing resistor (distances $D1$ and $D2$). This fact permits tuning over a relatively wide range by mechanically varying the position of the tunnel diode. One commercially available model using this method of tuning gives an output range from 1000 to 1500 mc. Electrical tuning over the desired frequency range can be achieved by placing a variable capacitor in a region of high electrical field. Similar experimental oscillators have been built to operate at 600 mc with 10 mw of output power and 7000 mc with 12 μw (microwatt) of output power. The power output is primarily dependent upon the diode peak current as well as the operating frequency.

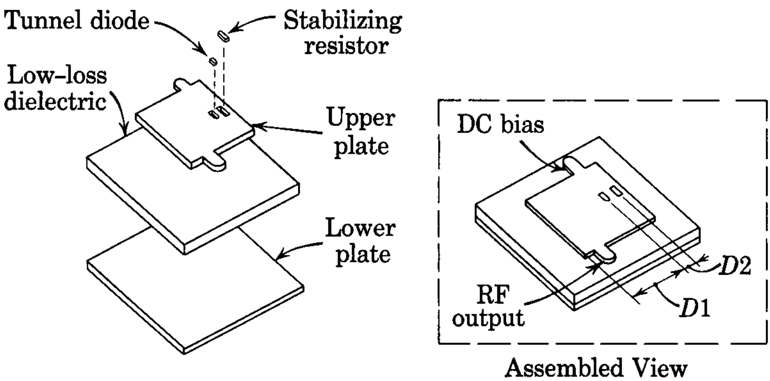


FIG. 7–11. Exploded view and assembled view of tunnel diode oscillator using strip transmission-line construction

7–9. Cylindrical Cavity (100 kmc) Oscillator

Research in tunnel diode applications at Bell Telephone Laboratories has resulted in cylindrical cavity oscillators operating at fundamental frequencies of 3 to 40 kmc. Detectable harmonic frequencies as high as 100 kmc have been produced. The cylindrical cavities (Fig. 7–12) act as tuned circuits, the resonant frequency depending upon the dimensions of the cavity. A tunnel diode was placed in each cavity at the apex (small end) of the re-entrant cone. The signal is coupled out of the circuit through a

loop from a coaxial cable placed at the large end of the cone. Resonance of the cavity occurs approximately at all frequencies for which the radius equals an odd integral number of quarter wavelengths. Odd harmonics through the seventh were produced for fundamental oscillations of approximately 3 kmc. The higher the fundamental oscillation, the lower the order of the odd harmonics detected. The microwave power output ranged from 50 μw at the lower frequencies to fractions of a microwatt at the higher frequencies. Continued research throughout the industry is expected to increase both the frequency limit and the available output power.

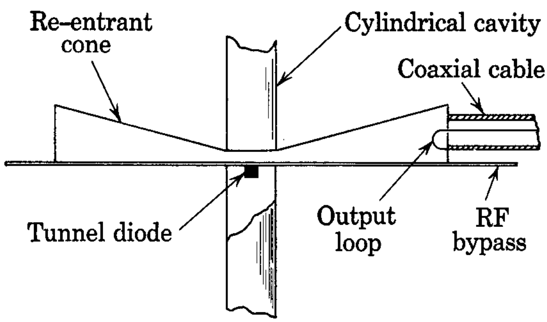


FIG. 7–12. Simplified version of a cylindrical cavity oscillator

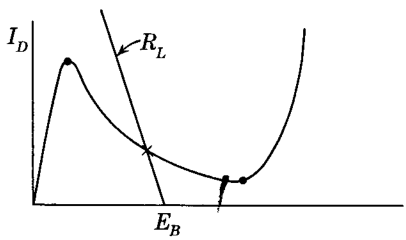
SECTION III. RELAXATION OSCILLATORS

7–10. Single Tunnel Diode, General

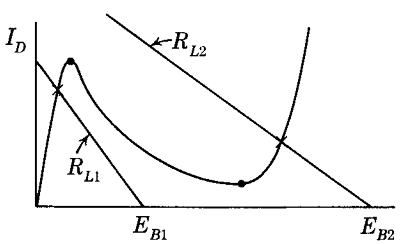
As previously indicated (par. 7–2) relaxation oscillators are classified as astable (*a* below), monostable (*b* below), or bistable (*c* below). The classification into which a particular relaxation oscillator falls depends upon the *value* of positive resistance in the circuit, the *magnitude* of dc voltage applied, and the current-voltage characteristic of the particular tunnel diode. For instance, the simplest tunnel diode relaxation oscillator is shown subsequently in Fig. 7–14A. Without knowledge of the value of resistance ($R1$), applied voltage (E_B), and the diode $CR1$ characteristic, it would be impossible to classify this circuit as an astable, monostable, or bistable relaxation oscillator. As a matter of fact, if inductance $L1$ is small enough to resonate effectively with the junction capacitance of the diode, and resistor $R1$ is smaller in value than the negative resistance of the diode, this circuit would be a sinewave oscillator.

a. Astable Circuit. Figure 7–13A shows the required relationship between the load resistor (R_L) value, the applied voltage (E_B), and current-

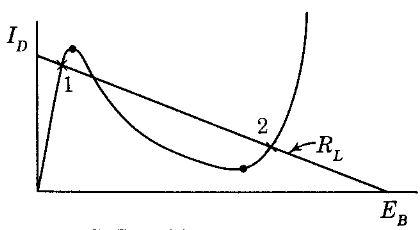
voltage characteristic of the diode. The load line must intersect the diode characteristic only within the negative-resistance region of the diode. The *slope* of the load line depends upon the load-resistor value only. However, the position of the load line depends upon the load-resistor value and the applied voltage. The load-resistor value must be *less* than the negative resistance of the diode, and the applied voltage must be greater than the diode peak voltage and less than the diode valley voltage (marked by dots). The point of intersection on the voltage axis is at the applied voltage point; it is assumed that at this point there is zero voltage across the load resistor. The point of intersection on the current axis is found by dividing the applied voltage by the load-resistor value; it is assumed that at this point there is zero voltage across the diode. Using this technique, then, the point of intersection (marked ×) of the load line and the diode characteristic divides the applied voltage (E_B) between the diode and the load resistor. From zero to the point of intersection indicates the voltage across the diode; from the point of intersection to E_B indicates the voltage across the load resistor. If no reactive (energy storing) elements, such as coil $L1$, were in the circuit (Fig. 7–14A), a steady-state dc current marked × (Fig. 7–13A) would flow continuously through the circuit. The reactive element in the circuit, however, continuously adds or subtracts to the applied voltage (E_B) and effectively moves the load line to the extreme right (beyond valley voltage) and the extreme left (below peak voltage) to cause a continuously varying voltage across each element in the circuit except the battery (par. 7–11). The rate at which this action occurs depends upon the time constant of the circuit (par. 7–2). A continuous out-



A. Astable (free-running)



B. Monostable



C. Bistable

FIG. 7–13. Relationships between load line and tunnel diode characteristic for astable, monostable, and bistable relaxation oscillators

put results, because the circuit is not stable (or is *astable*) or referred to as *free-running*.

b. Monostable. A monostable relaxation oscillator results if the load line intersects the diode characteristic at *one* point only; this point must be in one of the positive-resistance regions of the diode characteristic (Fig. 7–13B). The positive-resistance regions occur below the peak-current point and above the valley-current point. The monostable relaxation oscillator is not a free-running oscillator even though reactive, energy-storing elements, such as an inductance, may be included in the circuit. After the application of power to such a circuit, a steady-state condition results wherein the circuit comes to rest at the point of intersection between load line and diode characteristic. Since this point of intersection *does not occur* in the negative-resistance region where the diode is capable of shifting dc battery power to an ac form (Chapter 3), a stable condition occurs and a continuous steady current flows. The circuit remains in this condition until a disturbance (such as a voltage or a sinewave) is introduced from a separate source. This disturbance effectively increases or decreases the applied voltage.

1. Assume that the circuit has a load line R_{L1} as indicated; assume further that a pulse input raises the point of intersection to the peak-current point. At this point the circuit is unstable (negative resistance). Even though the original disturbances may no longer exist, the circuit will switch rapidly through the negative-resistance region until the point of intersection is in the high-voltage positive-resistance region of the diode. The extra voltage and power required to accomplish this switch is supplied to the circuit by the energy stored in the reactive component of the circuit. When this energy is dissipated (time dependent upon L/R or CR), the circuit *returns* to its *original* stable condition and remains at that point until other disturbances are introduced.

2. Assume that the circuit has a load line R_{L2} as indicated; assume further that a pulse input lowers the point of intersection to the valley-current point. At this point the circuit is unstable (negative resistance). Even though the original disturbance may no longer exist, the circuit will switch rapidly through the negative-resistance region until the point of intersection is in the low-voltage positive-resistance region of the diode. The extra voltage and power required to accomplish this is supplied to the circuit by the energy stored in the reactive component of the circuit. When this energy is dissipated, the circuit returns to its *original* stable condition and remains at that point until other disturbances are introduced.

3. A monostable relaxation oscillator, therefore, under steady-state con-

ditions will always be found in one particular point of bias current and diode voltage regardless of the previous history of the circuit.

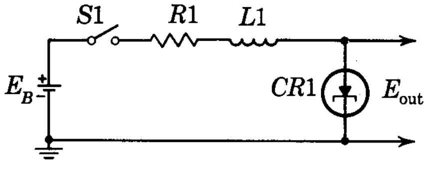
4. It is interesting to note that the circuit represented by Fig. 7–13A can be changed from astable to monostable by simply increasing the bias voltage above the valley-point voltage, or decreasing it below the peak-point voltage.

c. Bistable. A bistable relaxation oscillator results if the load line intersects the diode characteristic in both positive-resistance regions (Fig. 7–13C). The bistable circuit is not a free-running oscillator even though reactive elements may be included in the circuit. Assume that power is applied to the circuit represented. Current will increase to point 1 only; at this point a stable steady-state condition results with the applied voltage divided between the diode and the load resistor as indicated. If a disturbance such as a pulse voltage raises the current to the peak point, the circuit will switch rapidly to the high-voltage state. The circuit then comes to rest at point 2 which is also a steady-state condition. Note that at this steady-state point there is less current through the circuit with the applied voltage divided so that there is a greater voltage across the diode than across the load resistor. The circuit will remain in this condition until another disturbance is introduced to reduce the current to the valley-point value, and the circuit will switch to point 1. A bistable relaxation oscillator, therefore, has two points of steady-state values; the steady-state value at any particular time *depends upon the previous history of the circuit.*

7–11. Single Diode Multivibrator

The most common type of relaxation oscillator is the multivibrator. The electron-tube and the transistor versions of this circuit require two each of the particular active circuit device used. With only *one* tunnel diode a multivibrator circuit can be designed.

a. Figure 7–14A shows a multivibrator constructed with one load resistor ($R1$), coil $L1$, and diode $CR1$ connected in series to a bias voltage (E_B) through power switch $S1$. The relationship between the load line and the



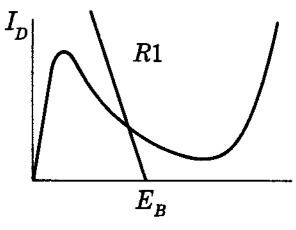
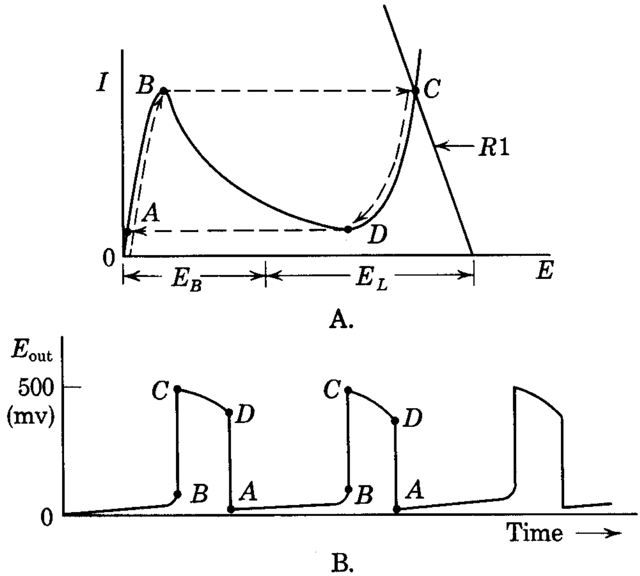


FIG. 7–14. Astable multivibrator, and load line and diode curve relationship



B.

FIG. 7–15. Tunnel diode characteristic showing switching action and output of astable multivibrator

diode characteristic (Fig. 7–14B) indicates that the circuit is astable (par. 7–10).

b. Assume that switch *S*1 is closed. The current rises in the circuit from zero to peak current, point *B* (Fig. 7–15A). The output voltage (across the diode) rises relatively slowly from zero to point *B* (Fig. 7–15B). At point *B*, the current would have a tendency to decrease; however, the action of coil *L*1 is to maintain a steady current flow in the same direction. The field about the coil starts to collapse and induces an instantaneous voltage (E_L) which adds to the bias voltage (E_B). Note that at this instant the load line intercepts the diode characteristic at point *C*, where the current through the diode equals the peak current but the diode is in the high-voltage state. This action happens almost with the speed of light, so the diode does not *pass* through the negative-resistance portion of the characteristic, but jumps from *B* to *C*. The rise in output voltage from *B* to *C* is shown in Fig. 7–15B. At point *C* the field about the coil starts to collapse (coil *L* discharges) and the current in the circuit falls from point *C* to point *D*. The drop in output voltage (*C* to *D*) is shown in Fig. 7–15B. At point *D* (valley-current point) the current would tend to rise; again the coil tends to maintain the same current. The diode then switches rapidly from point *D* to point *A*. The drop in output voltage from *D* to *A* is shown in Fig. 7–15B. This completes one cycle and a new cycle begins. Note that the voltage across the coil

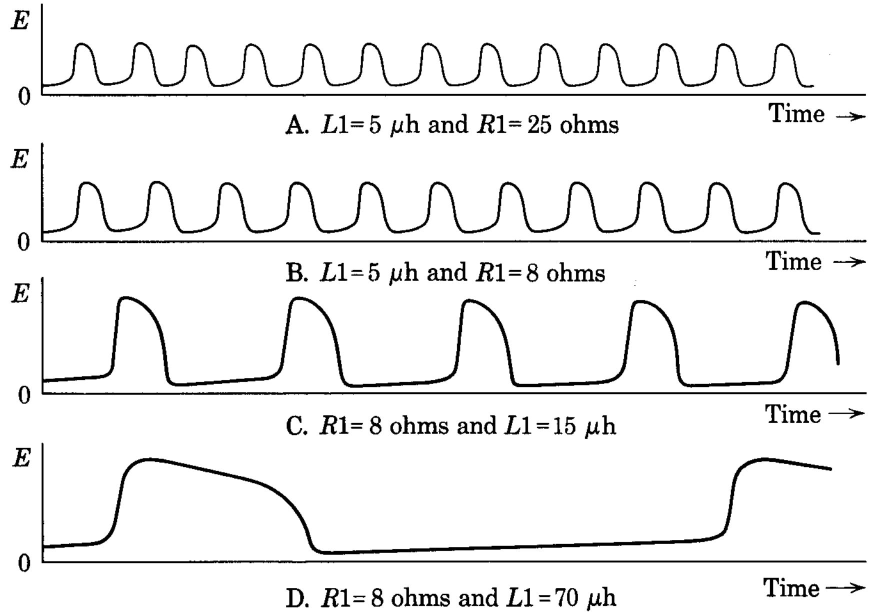


FIG. 7–16. Variations in output waveforms of multivibrator as resistance and inductance are varied

opposes the applied voltage from zero to B, *aids* the applied voltage from B to C, continues to aid the applied voltage from C to D but in diminishing quantities, and *opposes* the applied voltage from D to A.

c. A plot of the output voltage against time is shown in Fig. 7–15B. The periods represented by A–B and C–D are called *dwell* periods; the periods represented by B–C and D–A are called switching periods. Note that the dwell period (A–B) in the low-voltage region is much longer than the dwell period (C–D) in the high-voltage region. The dwell period is determined mainly by the charging and discharging of the coil (time constant) through the *total* resistance in the circuit. The time constant equals the ratio of inductance to resistance (L/R). During the A–B dwell period, the resistance of the diode is less than the resistance of the diode during the C–D dwell period. Note that the slope of the diode characteristic from A to B differs from the slope from C to D. This condition can be overcome with a single diode multivibrator by using a delay line (such as a long cable) in place of coil $L1$. Another method overcoming this problem uses a coupled pair (par. 7–12).

d. Variations of output waveforms with changes in resistance and inductance values are shown in Fig. 7–16. A 12-mc multivibrator resulted when a 5-μh coil and a 2-ohm resistor were used.

1. A and B show the decrease in frequency as the resistor is changed from 25 to 8 ohms, respectively. A decrease in resistance causes an increase in time constant (L/R).

2. C and D show the decrease in frequency as the inductance is changed to 15 and 70 μh, respectively. An increase in inductance causes an increase in time constant (L/R).

7–12. Coupled Pair, General

The development of the coupled-pair current-voltage characteristic was considered in paragraph 6–7. Like the single tunnel diode (par. 7–10), the coupled pair (Fig. 6–9) can also be used in astable, monostable, and bistable modes of operation. A basic coupled-pair relaxation oscillator is shown in Fig. 7–17. As in the case of the single diode, the mode of operation of this circuit depends upon the relationship of the current-voltage characteristic (at terminals 2–2′) and the *effective load line* caused by resistors $R1$ and $R2$ at terminals 2–2′.

a. Astable Operation. For astable operation the effective load line must intersect the current-voltage characteristic *only* in the negative-resistance region (Fig. 7–18A). It can be shown that if resistors $R1$ and $R2$ are equal, the load line will pass through the origin (i.e., $E = 0$ and $I = 0$). Furthermore, if resistor $R1$ and resistor $R2$ are each *less* than $2E_p'/I_p'$, the effective load line will intersect the current-voltage curve at only one point. Note that E_p' is the voltage at terminals 2–2′ when peak current I_p' passes through the terminals. Under these conditions, with an energy storing element (coil $L1$, Fig. 7–17) a free-running multivibrator can be constructed. Operation of such a circuit is discussed in paragraph 7–13.

b. Monostable Operation. For monostable operation the effective load line must intersect the current-voltage curve at only one point in either of the positive-resistance regions (Fig. 7–18B). Since the effective load line must not pass through the origin, resistors $R1$ and $R2$ cannot be equal (*a*

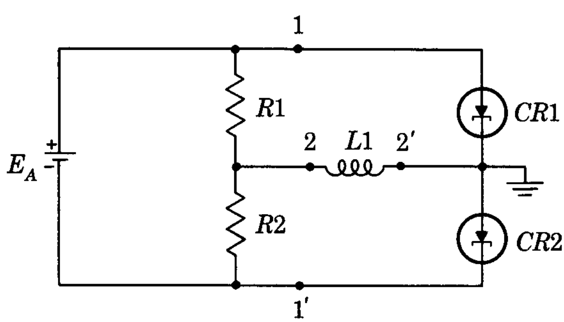


FIG. 7–17. Simplified coupled-pair relaxation oscillator

above). Their values must be chosen to establish the desired monostable point of operation. As in the case of the single diode when in a state of quiescence, the monostable coupled pair will always be found biased with the same voltage and current regardless of the previous history of the circuit.

c. Bistable Operation. Bistable operation requires the effective load line to intersect the current-voltage curve at one point in each positive-resistance region (Fig. 7–18C). If resistors $R1$ and $R2$ are equal, the effective load line will pass through the origin; in addition, if resistor $R1$ and resistor $R2$

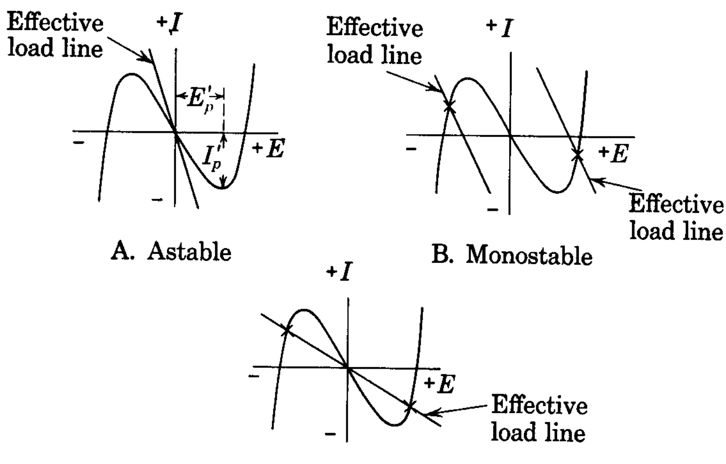


FIG. 7–18. Possible modes of operation of coupled-pair relaxation oscillator

are each made *greater* in value than $2E_p'/I_p'$, the effective load line will intersect the curve in each positive-resistance region. At quiescence the particular state of this circuit depends upon its previous history. Use of this mode of operation in switching circuits is covered in Chapter 8.

7–13. Coupled-Pair Multivibrator

a. Assume that the relaxation oscillator shown in Fig. 7–17 is so designed that resistors $R1$ and $R2$ form an effective load line that intersects the coupled-pair curve only at one point and in the negative-resistance region (Fig. 7–18A). The circuit constitutes a free-running multivibrator with an output (taken from terminal 2 to ground) as shown in Fig. 7–19B. Since diodes $CR1$ and $CR2$ have like current-voltage characteristics and resistors $R1$ and $R2$ are equal in value (par. 7–12*a*), one may conclude that each diode would draw an equal amount of current and no voltage would appear across coil $L1$; i.e., the circuit would act like a balanced bridge. In practice,

however, because of different tolerances, the diodes are not identical, nor are the resistors. Let it be assumed that when power is applied, the current through coil *L*1 (the difference in current between diodes *CR*1 and *CR*2) is the point *A* marked on the current-voltage curve (Fig. 7–19A). From *A* to *B*, the current continues to *change in the same direction* through coil *L*1; the corresponding points (*A–B*) are marked on the output waveform. The time it takes to go from *A* to *B* is directly proportional to the inductance of coil *L*1 and inversely proportional to the effective load resistance *and the resistance* (*slope*) *of the coupled-pair between A and B.* At *B* an unstable

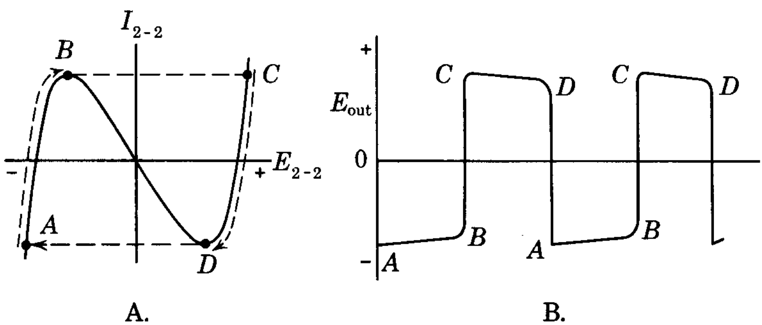


FIG. 7–19. Coupled-pair characteristic showing switching action and output of astable multivibrator

point is reached; the energy stored in the field of the coil discharges and switches the point of operation to *C* almost instantaneously. The current through the coil does not change in going from *B* to *C*. The corresponding points are marked on the output waveform. From *C* to *D* the current-charge through the coil is the same. The time duration from *C* to *D* depends upon the inductance of coil *L*1, the slope of the curve, and the effective load resistance as cited above. At point *D*, the coil again discharges, maintains a constant current, and switches almost instantaneously to point *A*. The cycle will repeat as long as power is applied. Compare the output waveform of the coupled pair (Fig. 7–19B) with the output of the single diode (Fig. 7–15B). With the coupled pair the duration of the dwell periods (*AB* and *CD*) are equal because the slopes of the coupled-pair curve in the positive-resistance regions are equal, whereas the slopes are unequal in the positive-resistance regions of the single diode.

b. A practical 10-mc coupled-pair multivibrator is shown in Fig. 7–20. Resistors *R*1, *R*2, *R*3, and *R*4 form voltage dividers which reduce the battery voltage to the desired value. The resistors also determine the effective load line for the coupled pair. The output of the multivibrator can be synchro-

nized to a sinewave or pulse signal coupled to the circuit through isolating resistor $R5$ and dc blocking capacitor $C1$.

7–14. Sawtooth Generator

a. A sawtooth generator using a tunnel diode ($CR1$) and a transistor ($Q1$) is shown in Fig. 7–21A. The composite characteristic of tunnel diode $CR1$ and transistor $Q1$, when connected in a circuit including resistors $R1$ and $R2$ as shown, is an N-type characteristic (par. 6–17). The transformation from an S-type characteristic for the diode to an N-type characteristic for the combination is discussed in paragraph 6–18.

b. Assume that power is applied to the circuit by closing switch $S1$. A sawtooth output will result (Fig. 7–21B). From 0 to A, capacitor $C1$ charges slowly through resistor R_s with current flow (solid-line arrow) as indicated. During this period transistor $Q1$ is off (nonconducting) and the voltage across diode $CR1$ is less than its peak-current voltage. At point A, diode $CR1$ reaches its peak-current voltage and switches rapidly to its high-voltage state (beyond valley-current voltage). This high voltage on the transistor $Q1$ base turns on transistor $Q1$. Capacitor $C1$ discharges rapidly (dashed-line arrows) through the low-resistance emitter-collector circuit. When the voltage across capacitor $C1$ falls to point B, the diode switches rapidly to its low-voltage state and turns off the transistor. The cycle then starts anew and continues as long as power is applied.

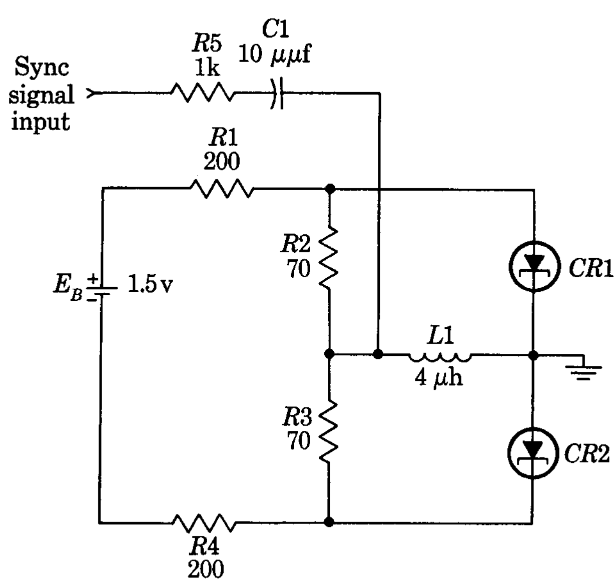


FIG. 7–20. Ten-mc coupled-pair multivibrator

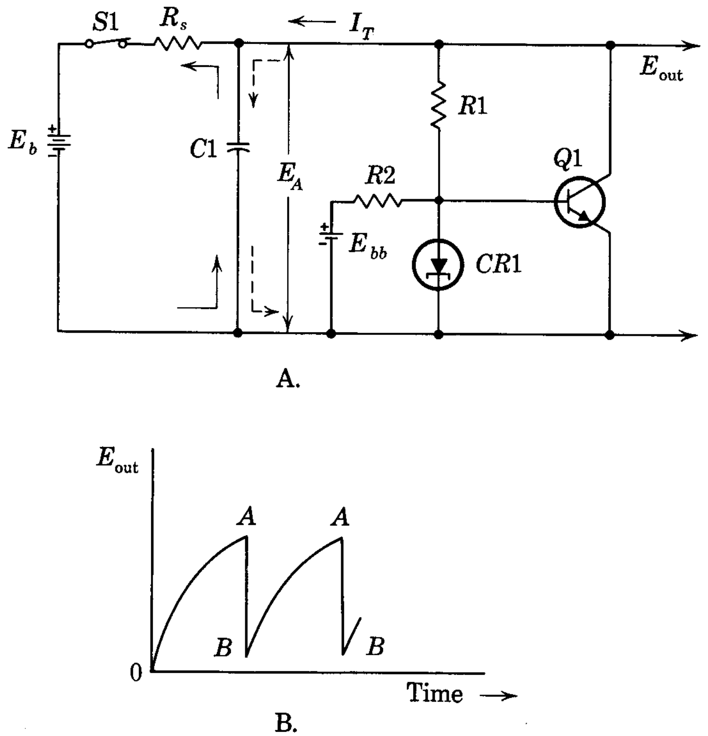


FIG. 7–21. Tunnel diode and transistor sawtooth generator

7–15. Diode-Transistor Multivibrator

An astable multivibrator formed by a tunnel diode and a transistor is shown in Fig. 7–22. The output voltage waveform is also shown. Resistor $R3$ is the collector load resistor. Resistors $R1$ and $R2$ with capacitor $C1$ form the main part of the timing circuit. When power is first applied, capacitor $C1$ charges through resistors $R2$ and $R3$. Diode $CR1$ is in its low-voltage state and transistor $Q1$ is cut off. While capacitor $C1$ is charging, the voltage at the collector almost equals the supply voltage (E_b). When the voltage across diode $CR1$ reaches the peak-current voltage, the diode switches rapidly to the high-voltage state and turns on transistor $Q1$ which immediately saturates and draws a heavy current. The collector voltage drops to a very low value and remains low until capacitor $C1$ discharges through resistor $R2$ and transistor $Q1$, as well as resistor $R1$, diode $CR1$, and the transistor base-emitter junction. When the voltage across diode $CR1$ falls to the valley voltage, diode $CR1$ switches to the low-voltage state and turns off transistor $Q1$. The voltage at the collector immediately rises to a value almost equal to the supply voltage. On the output waveform,

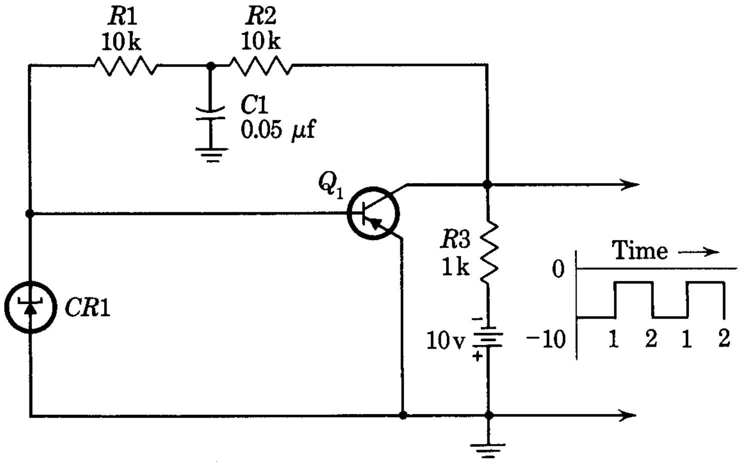


FIG. 7–22. Tunnel diode and transistor multivibrator

the time from 0 to 1 the transistor is *off;* from 1 to 2 the transistor is *on.* With the values of components indicated in the circuit, the output frequency is approximately 4 kc.

7–16. Summary

a. A tunnel diode can be made to oscillate below its self-resonant frequency by adding external inductance.

b. A tunnel diode can be made to oscillate above its self-resonant frequency by adding external capacitance.

c. A tunnel diode cannot be made to oscillate above its resistance cutoff frequency.

d. An oscillator formed by a diode in series with a reactive element has poor frequency stability.

e. An oscillator formed by a diode in series with an *LC* tank circuit has good frequency stability. If the frequency is low enough, a swamping capacitor in parallel with the diode further improves frequency stability.

f. Back diodes, which have high negative resistance, can be used in oscillators where high positive resistance exists because of high-resistance voltage dividers, or high-resistance coils.

g. High-resistance voltage dividers limit the drain on the bias battery.

h. At very low operating frequencies, the dc resistance of the heavy coils becomes appreciable.

i. The main disadvantage of the back diode is its very low power-handling capacity.

j. Crystal controlled tunnel diode oscillators take advantage of the crystal properties to maintain excellent frequency stability.

k. A voltage-tuned oscillator (900–1000 mc) may be formed with a tunnel diode and a portion of a coaxial transmission line (par. 7–7).

l. Strip transmission-line oscillators, operating in the 600 to 8000 mc range, can be mechanically or electrically tuned (par. 7–8).

m. Strip transmission-line construction is to microwave technology what printed circuits are to low-frequency technology.

n. Cylindrical cavity oscillators using tunnel diodes (par. 7–9) have produced usable (fractions of a microwatt) output signals as high as 100 kmc.

o. A free-running (astable) relaxation oscillator results if the load (resistor) line intercepts the tunnel diode characteristic in the negative-resistance region only.

p. A monostable relaxation oscillator results if the load (resistor) line intercepts the tunnel diode characteristic at one point *not* in the negative-resistance region.

q. A bistable relaxation oscillator results if the load line intercepts the tunnel diode characteristic in the low-voltage and the high-voltage positive-resistance regions.

r. An astable multivibrator can be formed using one tunnel diode; however, the low-voltage and high-voltage dwell periods are not equal in time duration.

s. A coupled-pair (tunnel diode) multivibrator produces a waveform that has equal dwell periods.

t. A sawtooth generator can be formed by using a tunnel diode and transistor whose composite characteristic is N-type.

u. A low-frequency multivibrator can be formed by using a tunnel diode, transistor, and an RC timing circuit.

Chapter 8

PULSE AND SWITCHING CIRCUITS

SECTION I. SWITCHING CHARACTERISTICS

8–1. General

a. Circuit Applications. Pulse and switching circuits are used in radar, television, telemetering, pulse-code communication, and computer equipments. The circuits operate as relaxation oscillators, amplifiers, inverters, frequency dividers, and wave shapers to perform the functions of limiting, triggering, gating, and signal routing. Some typical circuits are described in this chapter; triggered circuits are covered in Section II and gating circuits are covered in Section III.

b. Nonlinear Operation. Pulse and switching circuits are normally characterized by large-signal, or *nonlinear,* operation of the tunnel diode. These circuits usually require the application of a pulse for operation (par. 8–2). The input trigger pulses produce large and sudden changes in output voltage or output current. Nonlinear operation of this type usually results in output waveforms differing considerably from the input waveforms.

c. Unit Step Voltage. 1. Pulse waveforms widely encountered in large-signal operation of the tunnel diode are illustrated in Fig. 8–1. The *instantaneous* changes in voltage levels represent the *ideal* pulse. The effect of the tunnel diode on the ideal pulse is given in paragraph 8–4.

2. A voltage which experiences an instantaneous change in amplitude from one constant level to another is called a *unit step voltage.* When the unit step voltage is the *applied signal* to a switching circuit, it is usually of sufficient magnitude to cause the circuit output to change from a low-voltage state to a high-voltage state or vice versa.

3. Figure 8–1A shows a *positive* unit step voltage occurring at time $t1$. A positive unit step voltage increases a positive potential level to a higher positive potential level and decreases a negative potential level to a less negative potential level. Depending on the magnitude of the positive unit step voltage, a negative potential level may even be changed to a positive

potential level. Figure 8–1B shows a *negative* unit step voltage occurring at time $t2$. A negative unit step voltage increases a negative potential level to a higher negative potential level and decreases a positive potential level to a less positive potential level or even to a negative potential level. Figure 8–1C shows the formation of an ideal pulse (square or rectangular) by two unit step voltages, one positive (time $t1$) and one negative (time $t2$).

Note: Unit step currents occur in the same manner as unit step voltages. Current levels, rather than voltage levels, undergo instantaneous positive or negative changes.

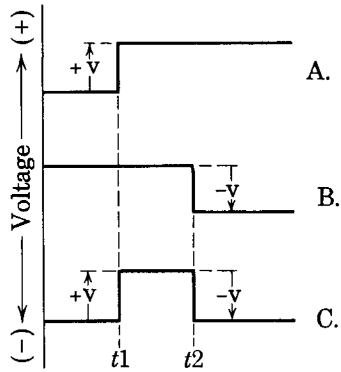


FIG. 8–1. Positive and negative unit step voltages, showing formation of a pulse

8–2. Types of Operation

All pulse and switching circuits are classified as astable, monostable, or bistable. These terms are discussed in detail in paragraph 7–10 as they apply to a single tunnel diode circuit, and in paragraph 7–12 as they apply to a coupled-pair circuit. In addition, astable (free-running) relaxation oscillators are covered in Chapter 7. Monostable and bistable circuits are covered in paragraphs 8–5 through 8–9. The outstanding aspect of each circuit is as follows:

a. The astable circuit requires only the application of dc power for operation. The values of the circuit elements will determine the rate at which the circuit will switch from a high-voltage level to a low-voltage level and vice versa. Trigger input pulses may be introduced into the circuit *only to synchronize the output of the circuit with another signal.*

b. The monostable circuit requires dc power for operation as well as an input trigger pulse to switch it from its normally stable condition in the high- or low-voltage state (positive-resistance regions). After the trigger pulse *initiates* the change in the state of the circuit, the circuit will of its own power switch to the nonstable state and then return to its single stable condition. The time required for each complete action depends upon the values of the circuit elements. The action is repeated for each input pulse.

c. The bistable circuit requires dc power as well as *two* input trigger pulses for a *complete* switching action. Depending on its previous history, the bistable circuit may be found at rest in either the low-voltage or the high-voltage state. One trigger pulse will initiate action to switch the circuit from the state in which it is *found* to the other state; a *second* trigger

pulse is required to initiate action to switch the circuit to its original state. The speed with which the circuit switches from one state to another depends primarily on the parameters of the particular tunnel diode used (par. 8–4).

8–3. On and Off States

The terms *on* and *off,* when used to describe electron tubes or transistors operating as switches, are for the most part self-explanatory. The term *on* with respect to these devices refers to heavy (usually saturation) current flow and low voltage between cathode and plate or emitter and collector, respectively. The term *off* refers to zero (or very low) current flow and high voltage between cathode and plate or emitter and collector. In the case of the tunnel diode (Fig. 8–2) the term *off* means that the tunnel diode is in the high-voltage state (biased beyond the valley-current point); the term *on* means that the tunnel diode is in the low-voltage state (biased below the peak-current point). In a particular application as a switch the tunnel diode current in the *off* or the *on* state may very well be the same magnitude; therefore these terms do not necessarily imply differences in *current* flow. The terms are arbitrarily chosen and used to mean just the opposite in some of the literature. The terms *on* and *off* as defined here, however, with respect to the tunnel diode keep the same *voltage* output condition consistent for all three devices (electron tube, transistor, and diode); i.e., *on* indicates low voltage and *off* indicates high voltage across the particular device.

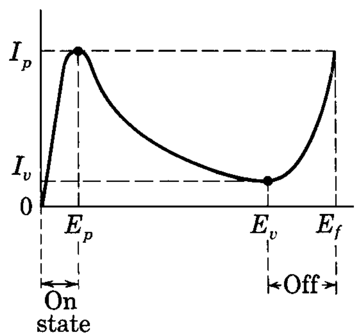


FIG. 8–2. Tunnel diode showing *on* and *off* switching regions

8–4. Tunnel Diode Switching Speed

a. The speed with which the average tunnel diode switches from the *on* to the *off* state or vice versa is very high. Some practical diodes switch in less than one nanosecond (mμs). The speed is determined mainly by the diode junction capacitance and the magnitude of the trigger pulse current. A current pulse which momentarily raises the diode current to a value larger than the peak current (Fig. 8–2) will switch the diode from the *on* state to the *off* state. A current pulse which momentarily lowers the diode current to a value less than the valley current will switch the diode from the *off* state to the *on* state. A measure of the switching speed is referred to as the *tunnel diode voltage rise time.* The rise time indicates the time

required for the pulsed tunnel diode voltage to climb from 10% of its maximum value to 90% of its maximum value. The formula used to calculate the rise time is as follows:

$$t_r = \frac{(E_f - E_p)C_D}{(I_p - I_v)10^3}$$

wherein:

t_r = rise time in nanoseconds

E_f = tunnel diode forward injection voltage in mv

E_p = peak-current voltage in mv

C_D = junction capacitance in $\mu\mu$f

I_p = peak current in ma

I_v = valley current in ma

This formula is based on a tunnel diode fed from a constant-current load line and triggered by a pulse having a minimum amplitude. An example of the use of this formula is given in *b* below.

b. Assume that it is desired to find the rise time of a tunnel diode having the following values:

E_f = 500 mv

E_p = 70 mv

C_D = 40 $\mu\mu$f

I_p = 10 ma

I_v = 1 ma

Substitute these values in the formula for rise time (*a* above):

$$t_r = \frac{(500 - 70)40}{(10 - 1)10^3}$$

$$= \frac{17{,}200}{9 \times 10^3}$$

Therefore, t_r = 1.9 nanoseconds.

Note: The values required to calculate t_r for a given tunnel diode are normally given in the manufacturer's specifications for the tunnel diode.

SECTION II. TRIGGERED CIRCUITS

8–5. Monostable Multivibrator, Single Diode

Figure 8–3A shows a monostable multivibrator, also referred to as a one-shot, single-shot, or single-swing multivibrator. The circuit is biased in the stable *on* region (*a* below) or the stable *off* region (*b* below). A single pulse causes the circuit to switch from its biased stable region to the other stable region and of its own accord returns to its biased stable region. The energy stored in the field of coil $L1$ furnishes the power required to switch the diode. Resistor $R1$ is the diode load resistor and its value in conjunc-

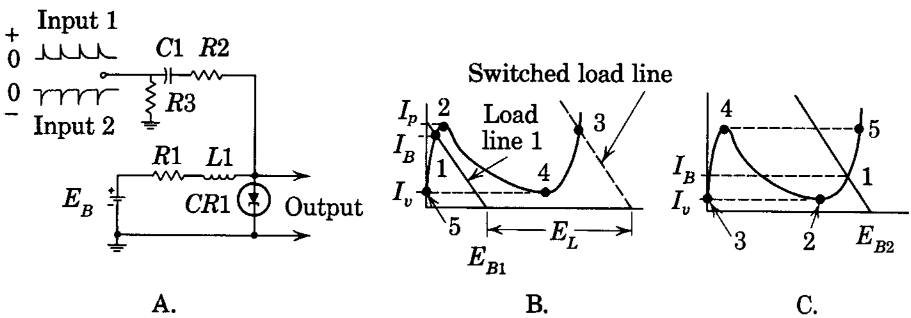


FIG. 8–3. Monostable multivibrator biased in the *on* or the *off* region

tion with the value of bias voltage (E_B) determines the stable region at quiescence. Capacitor $C1$ and resistors $R2$ and $R3$ form the pulse input network. Capacitor $C1$ is a blocking capacitor, resistor $R3$ is a dc return resistor, and resistor $R2$ is an isolating resistor that prevents loading of the pulse source by the tunnel diode.

a. On Region. Figure 8–3B shows the relation of the diode characteristic and the load line with a given bias voltage (E_{B1}). At quiescence the circuit is stable at point 1 and current I_B flows through resistor $R1$, coil $L1$, and diode $CR1$. Assume that a positive-going input pulse (input 1) of current equal in magnitude to $I_p - I_B$ is introduced through capacitor $C1$. The operating point rises to point 2, peak current. At this point when the current through the diode would normally fall (negative resistance), the field about coil $L1$ starts to collapse and switches the operating point almost instantaneously to point 3, also a stable point. Note that the action of coil $L1$ is to create a voltage (E_L) which adds to bias voltage E_{B1} and causes the load line to move to the right and intercept the diode curve at point 3; note also the switched load line. From point 3 to point 4 the field of coil $L1$ continues to discharge relatively slowly and voltage E_L is diminishing in magnitude. At point 4 (the valley current) the circuit again reaches the

negative-resistance (unstable) region where the current would tend to rise. The action of the coil is to keep the current constant and the circuit switches almost instantaneously to point 5. During this period the voltage across the coil opposes bias voltage E_{B1}. The circuit then returns relatively slowly to point 1, the biased stable condition. From point 5 to point 1 the opposing voltage of the coil diminishes to zero. The resultant output is one rectangular wave. Another positive-going pulse will cause the cycle to repeat.

b. Off Region. The same-valued circuit elements can be used to bias the multivibrator in the *off* region (Fig. 8–3C); only the bias voltage need be increased to that shown (E_{B2}). The circuit is now stable at quiescent point 1. A *negative*-going input pulse (input 2) is now required to switch the circuit through one complete cycle. The pulse magnitude must equal $I_B - I_v$. An action similar to that described in *a* above occurs with the circuit switching rapidly from point 2 to point 3, going relatively slowly from point 3 to point 4, rapidly from point 4 to point 5, and slowly from point 5 to point 1 to complete a cycle. The action will repeat with each negative-going pulse.

8–6. Bistable Multivibrator, Single Diode

a. The basic bistable single diode multivibrator is shown in Fig. 8–4A. It consists of a bias supply (E_B), a load resistor ($R1$), and a tunnel diode ($CR1$); capacitor $C1$ is used only to couple a trigger pulse into the circuit. An alternate method of introducing the trigger pulse is shown in Fig. 8–4B. Coil $L1$ has been added only to prevent shunting of the input pulse to ground by the bias battery. Note that the bistable circuit does *not* require an energy-storing element (an inductance) for operation, as in the case of the astable and monostable circuits. The discussion in *b* below applies to either circuit (Fig. 8–4A or B).

b. Figure 8–4C shows that the load line ($R1$) intersects the diode curve in two stable points, 1 and 2; the intersection in the negative-conductance region, of course, is unstable. A positive-going input pulse ($+E_p$) raises the current in the diode to the peak value, effectively increases the bias voltage ($E_B + E_p$), and causes the load line to rise so that it intersects the curve at 1′ and 2′. The diode switches almost instantaneously to point 2′; with the passing of the trigger pulse, the circuit comes to rest at point 2. Note that in going from point 1 to point 2 there has simply been a different division of the bias voltage (E_B); in the *on* state there was a small portion of the bias voltage across the diode and a high portion across the resistor; in the *off* state this condition is reversed. A negative-going input pulse ($-E_p$) lowers the current through the diode to the valley current, effectively lowers the bias voltage ($E_B - E_p$), and causes the load line to fall so that

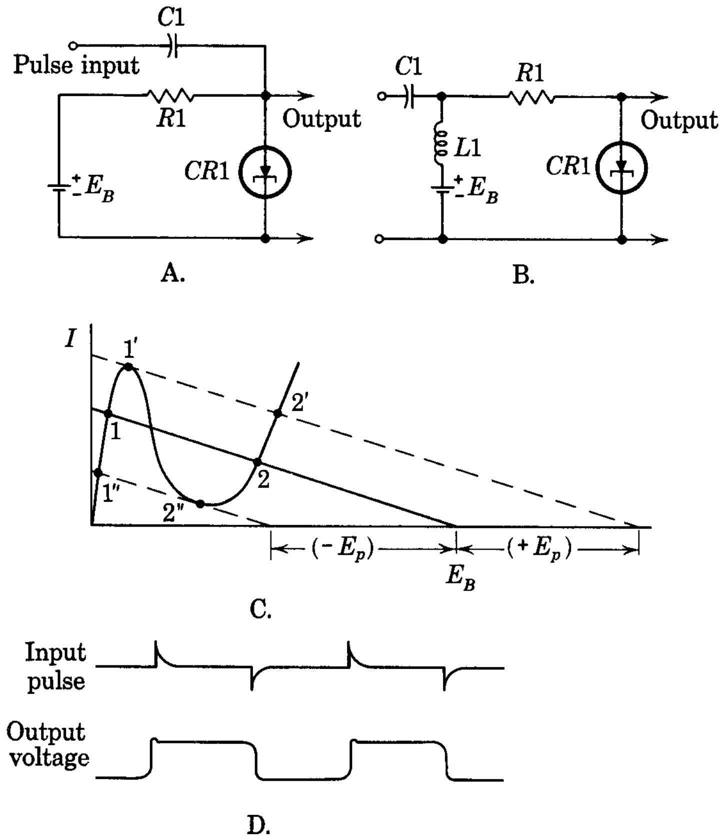


FIG. 8–4. Basic bistable circuit, showing alternate methods of triggering, load line on diode curve, and input pulses and output waveform

it intersects the diode curve at 1″ and 2″. The diode switches almost instantaneously to point 1″; with the passing of the trigger pulse the diode comes to rest at point 1. The relationship between the input pulses and the output waveform are shown in Fig. 8–4D.

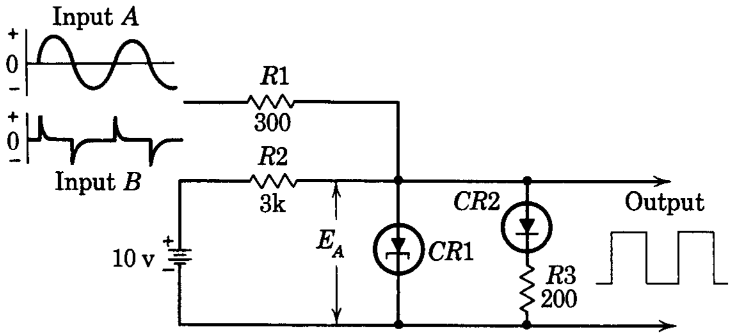
Note: If a negative input pulse is introduced when the circuit is in the *on* state (point 1), the load line is temporarily lowered and then returns to point 1 with no switching occurring. If a positive pulse is introduced when the circuit is in the *off* state (point 2), the load line is temporarily raised, but no switching occurs.

8–7. Bistable Multivibrator, Increased Sensitivity

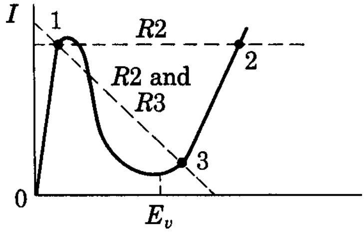
A very sensitive bistable multivibrator can be made by using a tunnel diode and a rectifying diode (*a* below), or by using a tunnel diode and a

transistor (*b* below). As used here, the sensitivity of a bistable multivibrator refers to the magnitude of the trigger pulses required to initiate switching action. The smaller the required pulse magnitude, the more sensitive the circuit.

a. Tunnel Diode and Rectifying Diode. High sensitivity can be obtained with the basic bistable multivibrator (Fig. 8–4A) by selecting a load resistor and a bias voltage so that the load line intersects the *on* region of the diode curve close to the peak current, and also intersects the *off* region of the diode curve close to the valley current. Such a load line is shown intersecting points 1 and 3 in Fig. 8–5B. Low-valued positive and negative trigger pulse currents would initiate switching resulting in high sensitivity. However, because of the difference in the slopes of the diode curves in the *on* and *off* regions, a nonsymmetrical output will result. The resistance in the positive region is higher than that in the *on* region. A circuit used to overcome this difficulty and still retain high sensitivity is shown in Fig. 8–5A. The battery, resistor $R2$, and tunnel diode $CR1$ form the basic bistable multivibrator (par. 8–6). Isolating resistor $R1$ prevents loading of the pulse source by the tunnel diode. To the basic circuit has been added a



A.



Note: Diode $CR2$ conducts at point 2, not at point 1

B.

FIG. 8–5. High sensitivity bistable multivibrator showing biasing technique

normal rectifying diode ($CR2$) and resistor $R3$. The normal rectifying diode, unlike the tunnel diode, does not start conducting when forward biased until the voltage is approximately equal to the tunnel diode valley voltage (E_v), and then it conducts heavily. Figure 8–5B shows the load line for resistor $R2$ and indicates that in the *on* state the circuit is biased at point 1; diode $CR2$ is nonconducting. A small, positive-going input pulse would switch the circuit rapidly to point 2. At point 2 diode $CR2$ conducts and effectively places low-valued resistor $R3$ in parallel with the tunnel diode. The effective load line composed of resistors $R2$ and $R3$ causes

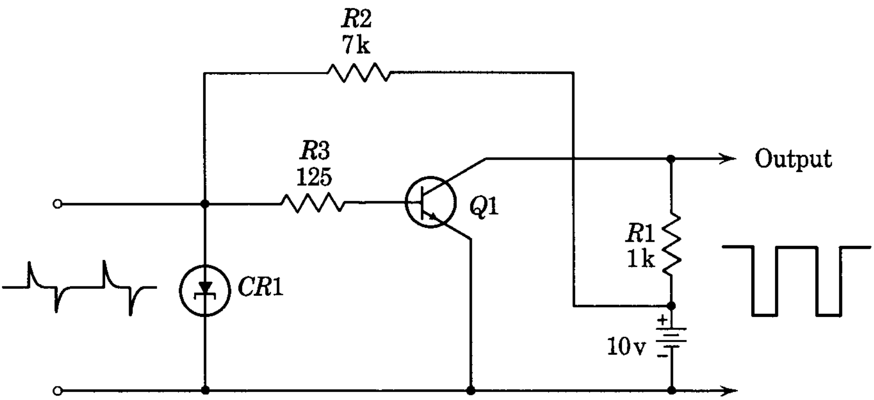


FIG. 8–6. Bistable multivibrator using tunnel diode and transistor

the circuit to be biased at point 3 in the *off* state. Because of the low resistance, the circuit falls rapidly from point 2 to point 3. A negative-going input pulse will switch the circuit rapidly from 3 to 1, causing diode $CR2$ to become nonconducting and setting the circuit for another cycle. The input to this circuit could be a sinewave (input A) or pulses (input B). If a sinewave is used, the circuit is referred to as a *squarer*.

b. Tunnel Diode and Transistor. A very sensitive bistable multivibrator using a tunnel diode ($CR1$) and an *n-p-n* transistor ($Q1$) is shown in Fig. 8–6. The main difference between this circuit and that shown in Fig. 8–5A is that rectifying diode $CR2$ has been replaced by the base-emitter junction of transistor $Q1$. When the tunnel diode is in the *on* state, resistor $R2$ biases the diode close to the peak current; the base-emitter junction of the transistor is nonconducting. When the tunnel diode is in the *off* state, the base-emitter junction conducts and parallels the tunnel diode with low-valued resistor $R3$ so that the circuit is biased close to the valley-current point of the tunnel diode. Resistor $R1$ acts as the collector load for transistor $Q1$. With the tunnel diode in the *on* state, the transistor is cut off and its

collector voltage is high (almost equal to the battery voltage). A small positive input pulse rapidly switches the tunnel diode to the *off* state; transistor $Q1$ conducts and saturates; collector voltage drops to near zero. A small negative input pulse rapidly switches the tunnel diode to the *on* state; transistor $Q1$ cuts off and its collector voltage rises to the battery voltage. The cycle is repeated with each pair of positive and negative pulses.

Note: If a *p-n-p* transistor is used, the battery and the tunnel diode must be reversed in their respective positions. A negative pulse will then switch the tunnel diode from the *on* to the *off* state; a positive pulse will switch the tunnel diode from the *off* to the *on* state.

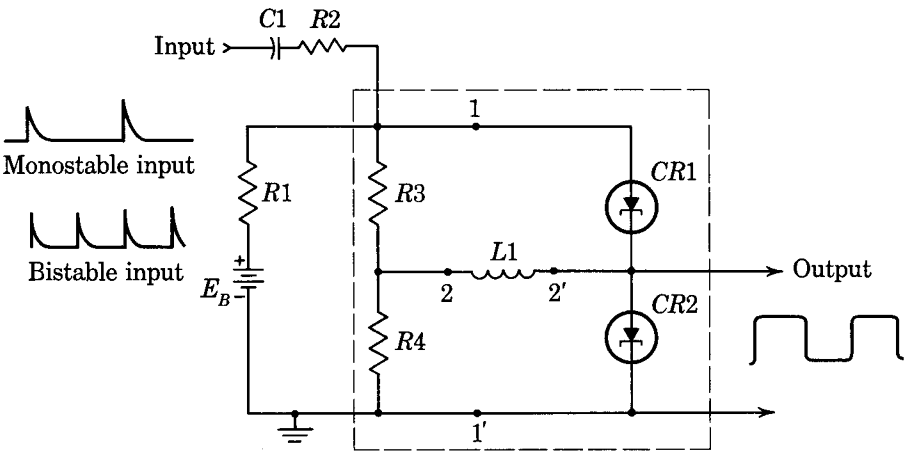


FIG. 8–7. Monostable or bistable coupled-pair multivibrator

8–8. Monostable or Bistable Multivibrator, Coupled Pair

The basic coupled-pair tunnel diode circuit is shown in dashed lines in Fig. 8–7A. The development of a family of characteristic curves for this *device* is covered in paragraph 6–7. An astable multivibrator using the coupled pair is covered in paragraphs 7–12 and 7–13. A monostable multivibrator is covered in *a* below; a bistable multivibrator is covered in *b* below. In the circuit shown, resistors $R1$, $R3$, and $R4$ form a voltage divider and establish the quiescent bias voltage for diodes $CR1$ and $CR2$. Capacitor $C1$ is a blocking capacitor and couples the trigger pulses into the circuit. Isolating resistor $R2$ prevents loading of the pulse source. Coil $L1$ is the energy-storing element required for switching action.

a. Monostable Circuit. For monostable operation, it is required that the values of resistors $R1$, $R3$, and $R4$ be so chosen that their combined *effective* resistance (actually in series with coil $L1$) results in a load line that inter-

sects the diode-pair curve at one point in a positive-resistance region. To accomplish this, resistors $R3$ and $R4$ must be different in value. Depending on the choice of values, load line 1 (Fig. 8–8A) may be chosen, or load line 2; in each case monostable operation will result. If load line 1 or

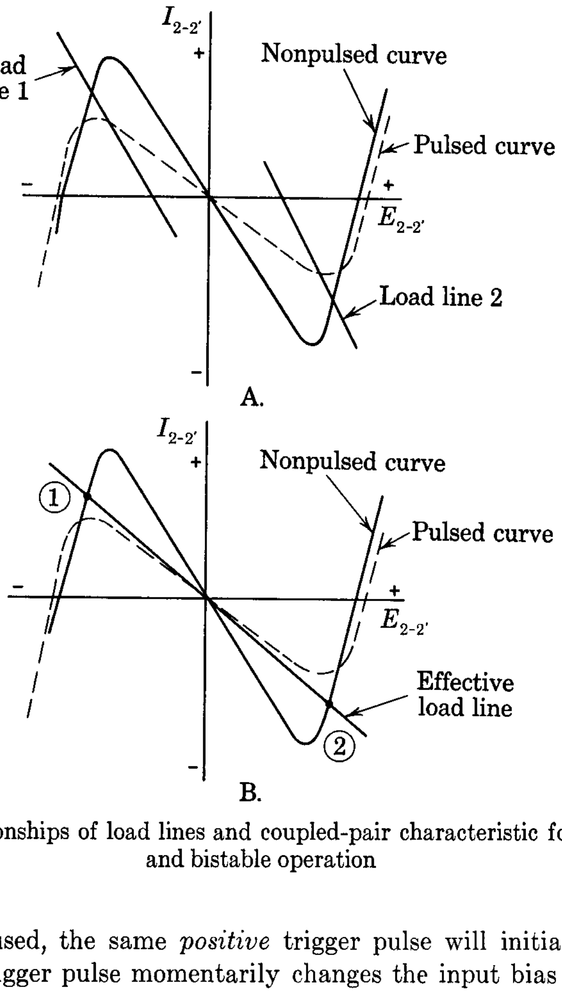


FIG. 8–8. Relationships of load lines and coupled-pair characteristic for monostable and bistable operation

load line 2 is used, the same *positive* trigger pulse will initiate switching action. The trigger pulse momentarily changes the input bias voltage and the diode-pair curve to that shown in dashed lines. This change causes the load line (either 1 or 2) to fall in the negative-resistance (unstable) region and rapid switching occurs. The load line moves to the opposite stable region and back to the *original* stable bias region to complete one cycle. This action occurs for each positive input pulse. Because the positive-resistance regions of the diode-pair curve are symmetrical, a symmetrical

output wave results. Note that the same action would occur if the trigger pulse were introduced in series with battery E_B.

b. Bistable Circuit. For bistable operation it is required that the load line intersect the diode-pair curve in each positive-resistance region (Fig. 8–8B). Such a load line will result if resistors $R3$ and $R4$ (Fig. 8–7) are equal in value and of the proper magnitude. A positive trigger pulse momentarily changes the input bias voltage and the diode-pair curve to that shown in dashed lines. This change causes the load line to fall into the negative-resistance (unstable) region and switching occurs. If the circuit was at rest at point 1, it will now come to rest at point 2; if it was at rest at point 2, it will now come to rest at point 1. Note that a *positive* input pulse can switch the circuit in *either* direction. A second positive trigger pulse will complete one cycle. In addition to having a *symmetrical* output waveform, the coupled-pair bistable multivibrator has a second advantage over the single tunnel diode bistable multivibrator (par. 8–6) in that triggering pulses of only *one polarity* are required by the coupled pair.

8–9. Controlled-Negative-Resistance Multivibrator

a. A two-tunnel diode controlled-negative-resistance *device* is shown in dashed lines in Fig. 8–9A. A discussion of the development of the family of characteristics for this device is given in paragraph 6–8. The most important property of this circuit is the variation in negative resistance at terminals 2–2′ in accordance with the control current introduced at terminals 1–1′ (Fig. 8–9B). This feature permits the construction of a multivibrator with an output waveform, the amplitude of which can be controlled. The circuit shown is a monostable multivibrator. The basic multivibrator is formed by load resistor $R3$, the energy-storing element (coil $L1$), and bias battery E_B in conjunction with the negative resistance displayed at terminals 2–2′. The values of load resistor $R3$ and battery voltage E_B are selected so that the circuit is stable only at one point in the *on* region of the *device.* Capacitor $C2$ couples the trigger pulses to the multivibrator. Resistor $R2$ is large in value and, with battery E_c, produces a constant-current source for biasing the input port. (Actually this bias selects the particular curve of the family of characteristics at which the device will operate at quiescence.) Capacitor $C1$ couples an input signal to the circuit to vary the control current as desired. Note that a step-voltage (staircase) signal is used here.

b. Each input trigger pulse causes the load line to rise above the peak current of the device and enter the unstable region; the circuit switches rapidly to the *off* state. In the *off* state the load line can intercept any of the device curves, depending upon the control current present at that instant.

After a period depending upon the values of resistor $R3$, and coil $L1$, and the resistance of the device in the *off* region, the circuit switches rapidly to its original bias point and completes one cycle. Each trigger pulse will repeat the cycle. As indicated in Fig. 8–9A the trigger pulses and the con-

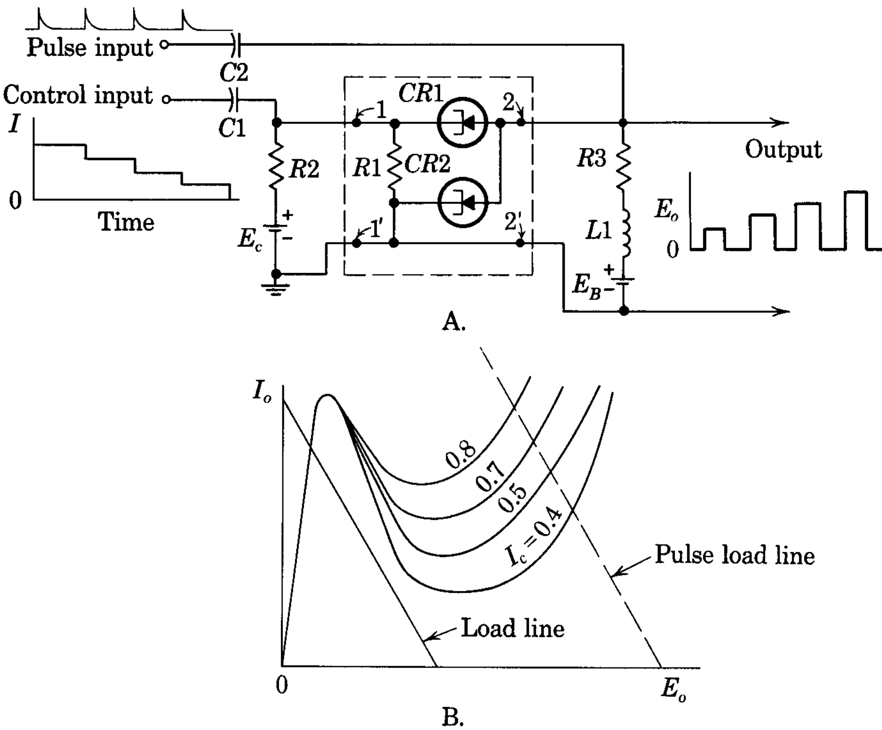


FIG. 8–9. Variable output monostable multivibrator showing relationship of load line and characteristic curves

trol current input signal are synchronized to produce the varying amplitude output.

c. The monostable multivibrator can also be biased in the *off* region of the *device.* In that case a negative trigger pulse would be required for switching. The negative trigger pulse would have to vary in magnitude as required by the control current input signal, or a large enough value would have to be used to ensure switching at any level of control current.

d. With proper biasing this circuit could also be used for astable or bistable operation. In addition, the three tunnel diode controlled-negative-resistance *device* could be used (par. 6–9).

8–10. Pulse Frequency Divider and Staircase Wave Generator

a. The circuit shown in Fig. 8–10 can be used as a pulse frequency divider or a staircase wave generator; these functions are performed simultaneously. In the circuit shown four tunnel diodes are connected in series. This number of diodes will produce a four-step staircase wave (not counting the step at quiescence) for each four input trigger pulses. After the fourth step the output returns to its original voltage level. In addition, the *fourth* input

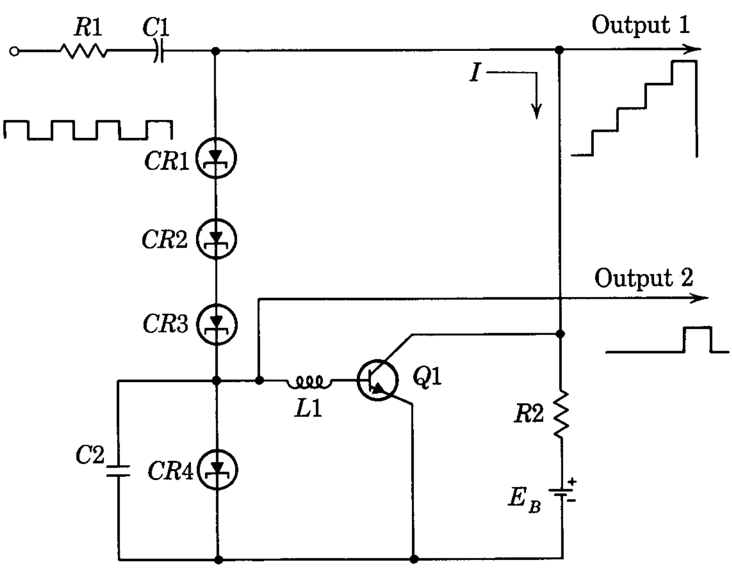


FIG. 8–10. Four-to-one pulse frequency divider and staircase wave generator

trigger pulse produces a pulse output across diode *CR*4, thus providing a four-to-one pulse frequency divider. A larger or a smaller number of diodes can be used; the number of steps of the staircase wave and the pulse frequency division ratio will correspond to the number of diodes used.

b. The development of the combined characteristic curve of series-connected tunnel diodes to form a multistate switch is covered in paragraph 6–5. Diode *CR*1 has the lowest peak current; each diode has a slightly higher peak current than the lower-numbered diode. Diode *CR*4, having the highest peak current, will switch last. Inasmuch as resistor *R*2 and battery E_B form a constant-current source that limits the voltage under quiescent conditions, all the diodes are initially in the *on* state. With diode *CR*4 in the *on* (low-voltage) state, transistor *Q*1 does not conduct. Resistor *R*2 also acts as the transistor collector load. Resistor *R*1 prevents loading of

the pulse source by the tunnel diode circuit. Capacitor $C1$ couples the trigger pulses into the circuit.

c. The first input pulse switches diode $CR1$ to the *off* state and raises the output voltage level by one step. The second, third, and fourth pulses switch diodes $CR2$, $CR3$, and $CR4$, respectively, to the *off* state; with each switching action the output voltage level is increased by one step. Immediately after diode $CR4$ switches, transistor $Q1$ conducts and saturates; its collector voltage drops to near zero and causes all the diodes to *reset* to the *on* state. This action in turn cuts off transistor $Q1$ collector current. A second set of pulses will repeat the cycle. Coil $L1$ and capacitor $C2$ isolate diode $CR4$ from the base-emitter circuit of transistor $Q1$. The time factor introduced by the two elements permits diode $CR4$ to switch to the *off* state before transistor $Q1$ conducts. Switching speed is thereby increased and efficient use is made of the trigger pulse input current.

d. The speed of the circuit is limited by the speed of the reset transistor. The rise time of the staircase wave depends upon the rise time of the trigger pulses. Note that operation of the circuit does not require trigger pulses equally spaced in time.

SECTION III. GATING CIRCUITS

8–11. General

Gating circuits are used most extensively in computers and computer-type equipment. The gating circuits discussed in this section are multiple input bistable tunnel diode circuits which will switch from one stable state to another stable state (produce an output) when certain conditions are met by the input signals. The input signals are most often pulses, or unit step voltages and currents. The conditions to be met by the input signals are usually implied by the designation assigned to a particular type gate.

a. The AND gate, also known as the COINCIDENCE gate, produces an output only if all the input terminals receive input signal pulses simultaneously (par. 8–12, 8–14, 8–15, 8–16, 8–17, and 8–21).

b. The NOT AND gate input requirements are the same as those for the AND gate (*a* above) except that the resultant output is opposite in polarity to the input signal pulses (par. 8–13). The NOT refers to phase inversion.

c. The OR gate produces an output if any one input is present *or* if all inputs are present simultaneously (par. 8–12, 8–14, 8–15, 8–17, and 8–21).

d. The NOR gate is identical with the OR gate except that phase inversion occurs (par. 8–13).

e. The NOT gate is actually a phase inverter (par. 8–18). Unlike the other gating circuits, multiple input signals are not required. A positive

unit step or a negative unit step input voltage results in a negative unit step or a positive unit step output voltage, respectively.

f. The EXCLUSIVE OR gate (par. 8–18) produces a high-level output if one input signal is high, and it will produce a low-level output if both input signals are high.

g. The MAJORITY gate (par. 8–19 and 8–20) has an *odd* number of input terminals. The polarity or magnitude of the output signal depends upon the predominant polarity of the input signals. Assume that *three* input pulses are present. If the three are not alike in polarity, the output polarity or magnitude is determined by the two pulses that are alike.

8–12. Biasing for AND or OR Gate

Figure 8–11A shows a basic bistable circuit with multiple inputs. Depending on the bias conditions established, the circuit can operate as an AND gate (*a* below) or an OR gate (*b* below). Resistors $R1$, $R2$, and $R3$ isolate the input pulse sources from each other and prevent loading of the pulse sources by the tunnel diode circuit. Resistor $R4$, in conjunction with bias battery E_B, determines the bistable operating points on the diode characteristic.

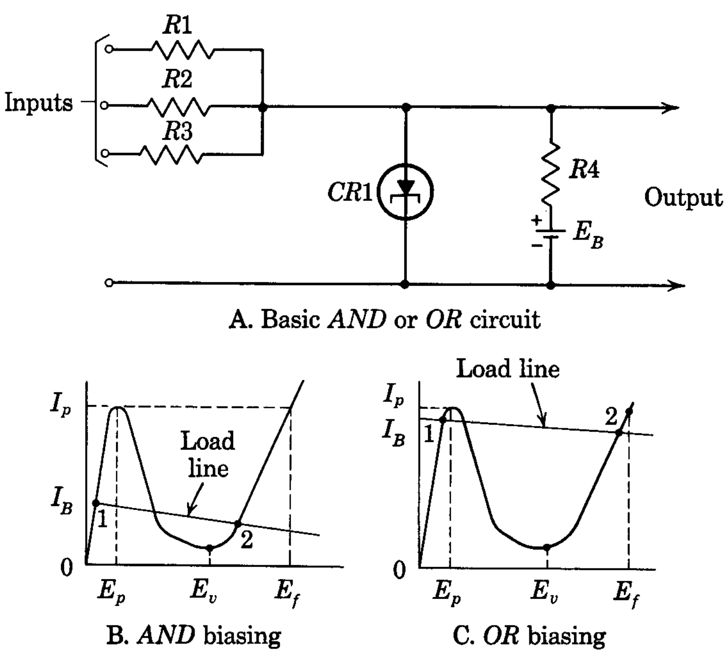


FIG. 8–11. AND or OR gate and biasing condition for each

a. AND Gate. Figure 8–11B shows the normal bias conditions for AND gate operation. Resistor $R4$ load line intersects the diode curve at *a very low point in the on region.* Initially the current through the diode at point 1 equals I_B. The total current input of the pulses required for switching to the *off* state (point 2) must be at least equal to $I_p - I_B$. AND gate operation is assured if all the input pulses are so restricted in magnitude that *all* must be present simultaneously to cause a current increase equal to $I_p - I_B$ to flow through the diode. When this condition is met, the circuit will switch to point 2, another stable point. To perform the AND function again, the circuit must be switched back to point 1. This condition can be achieved by reducing the bias voltage to zero, changing the bias resistor ($R1$) value, or introducing a negative pulse sufficient in magnitude to drive the circuit to the valley-current point causing switching to occur. The latter method is most often used. The pulse used to accomplish this action is referred to as a reset or clearing pulse.

b. OR Gate. Figure 8–11C shows the normal bias conditions for OR gate operation. Resistor $R4$ intersects the diode curve at a point close to the peak current in the *on* region. Initially the current through the diode equals I_B. Switching occurs if the current increase in the diode equals $I_p - I_B$. OR gate operation is assured if each input pulse signal causes a current increase equal to $I_p - I_B$ to flow through the diode. When this condition is met, the diode will switch to point 2 when one or more of the prescribed inputs are present. To repeat the OR function, a negative input reset pulse is required.

8–13. NOT AND or NOR Gate

Biasing techniques for the NOT AND or the NOR gate are identical with those for the AND or the OR gate, respectively (par. 8–11). Figure 8–12 shows a circuit that can be used for a NOT AND or a NOR gate. Circuit

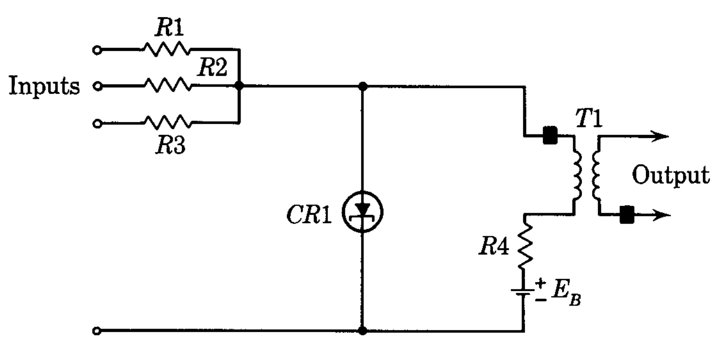


FIG. 8–12. NOT AND or NOR gate

elements in this circuit perform the same function as the correspondingly referenced circuit elements in Fig. 8–11 and have the same values for the corresponding circuit application. In Fig. 8–12, however, transformer $T1$ has been added to provide the negation feature, which is a phase-inversion at the output. In this manner an AND gate is transformed to a NOT AND gate, and an OR gate is transformed to a NOR gate.

8–14. Unidirectional AND or OR Gate

A circuit which can be used as an AND or an OR gate is shown in Fig. 8–13. The basic bistable circuit is inclosed in dashed lines and consists of load resistor $R5$, bias battery E_{B2}, and tunnel diode $CR5$. The circuit is fed from three previous stages through terminals 1, 2, and 3. One preceding stage (tunnel diode $CR1$, load resistor R_1, and bias battery E_{B1}) is shown. Resistors $R2$, $R3$, and $R4$ prevent loading of the input stages by diode $CR5$. *Backward* diodes $CR2$, $CR3$, and $CR4$ ensure unidirectional (left-to-right) flow of energy from the input stage to that in dashed lines. The stage under consideration feeds three succeeding stages connected to terminals 4, 5, and 6. One output stage (load resistor $R9$, tunnel diode $CR9$, and bias battery E_{B3}) is shown. Resistors $R6$, $R7$, and $R8$ prevent loading of diode $CR5$ by the succeeding stages. *Backward* diodes $CR6$, $CR7$, and $CR8$ ensure unidirectional (left-to-right) flow of energy from diode $CR5$ to the succeeding stages. Backward diodes (par. 6–2) must be used for the indicated purpose because the tunnel diode is a one-port, bidirectional device that uses the same two terminals for input and output. The operation of the backward diodes in performing their function is discussed below. Analysis of the circuit as an AND or OR gate is discussed in paragraph 8–15.

a. A comparison of a tunnel diode current-voltage curve and a backward diode current-voltage curve is made in Fig. 8–14A. When the tunnel diode goes through its peak current and valley current, negligible current (considered *leakage* current) flows through the backward diode (although it is forward biased). With reverse bias (less in *magnitude* than the tunnel diode peak voltage) the backward diode conducts very heavily. Because of this the backward diode can be used as a *rectifying* device in the necessarily *low-voltage* circuits of tunnel diodes. The high-voltage requirements of the normal (lightly doped) rectifying diode precludes its use in simple tunnel diode circuits. A comparison of the current-voltage curves of the backward diode and the normal rectifying diode is made in Fig. 8–14B. Note that the backward-diode curve has been reversed and inverted for comparison purposes. When *reverse biased,* the backward diode displays low resistance, whereas the normal rectifying diode displays low resistance when *forward biased* beyond the valley-voltage point of a tunnel diode made of the same material.

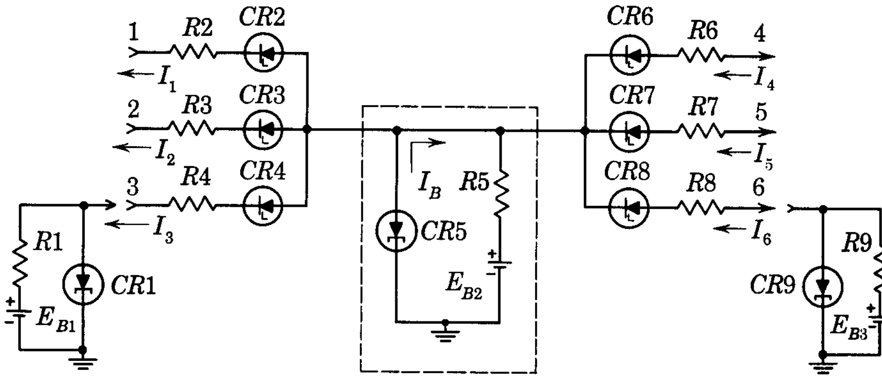


FIG. 8–13. AND or OR gate with multiple input and output terminals

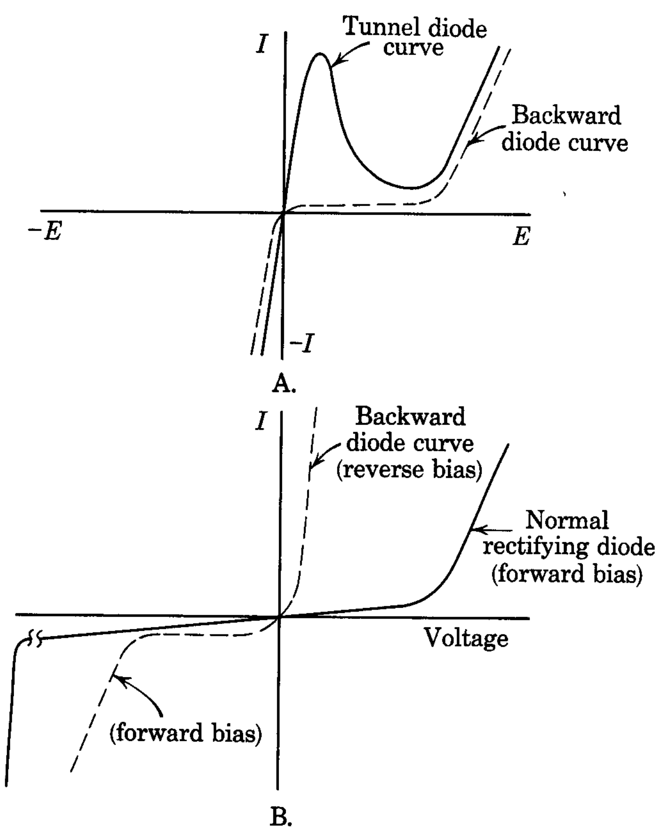


FIG. 8–14. Comparison of tunnel diode and normal rectifying diode current-voltage curves with backward diode current-voltage curve

b. The method by which the backward diode ensures unidirectional energy flow in the gate circuit (Fig. 8–13) is by acting as a rectifier within low-voltage ranges.

1. Figure 8–15A shows a partial schematic of the gate circuit. The input tunnel diode (*CR*1), one isolating resistor (*R*4), one backward diode (*CR*4), and the gate diode (*CR*5) are shown. When input diode *CR*1 is in the low-voltage state and gate diode *CR*5 is in the high-voltage state, backward diode *CR*4 is forward biased in the leakage current region which is a high-resistance region; this point of bias is marked by a dot on the current-voltage curve. The direction of electron flow representing the negligible leakage current is shown by a dashed-line arrow. The backward diode acts on an *open* switch and no energy flows right to left (from diode *CR*5 to diode *CR*1).

2. With the voltage conditions reversed for diodes *CR*1 and *CR*5 (Fig. 8–15B), backward diode *CR*4 is reverse biased (indicated by a dot on the current-voltage curve), and conducts heavily; it acts as a low-resistance or short circuit. The direction of electron-current flow is shown by the arrow. With diode *CR*1 at a higher voltage than diode *CR*5, *energy* flows from left to right, which is desirable.

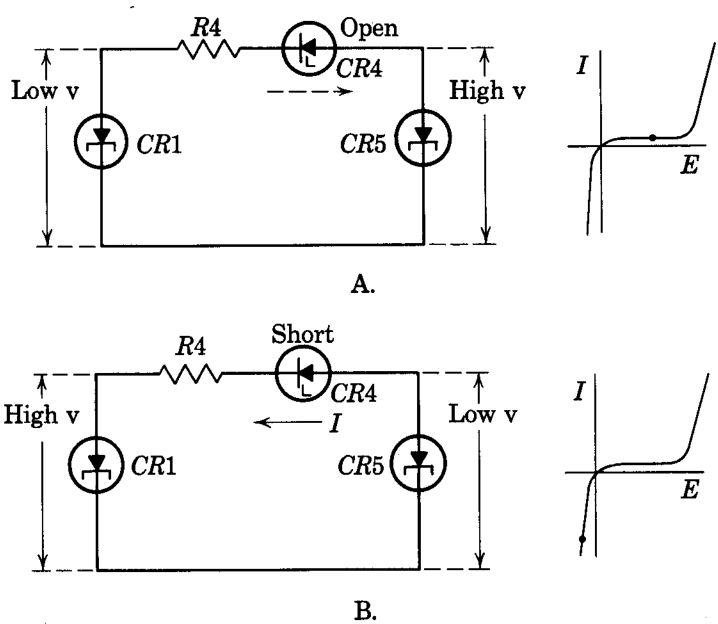


FIG. 8–15. Partial schematic, input and gate diodes, showing effect of different voltage states on backward diode

3. When diodes $CR1$ and $CR5$ are both in the high-voltage state or both in the low-voltage state, zero voltage occurs across backward diode $CR4$; there is no exchange of energy in either direction.

c. Figure 8–16A shows a partial schematic of the gate circuit. The gate diode ($CR5$), one backward diode ($CR8$), one isolating resistor ($R8$), and one output tunnel diode ($CR9$) are shown. With diode $CR5$ in the low-voltage state and diode $CR9$ in the high-voltage state, backward diode $CR8$ is forward biased in the leakage current region (marked by a dot) and acts as an open circuit. There is no flow of energy from right to left. With diode $CR5$ in the high-voltage state and diode $CR9$ in the low-voltage state, diode $CR8$ is reverse biased in the high-current region (marked by a dot); electron current flows in the direction shown by the arrow. Electrical *energy* flows from the high voltage to the low voltage (left to right). When diodes $CR5$ and $CR9$ are in the same voltage state, zero bias occurs across diode $CR8$ and no current or energy flows in either direction.

d. The backward diodes ($CR2$, $CR3$, and $CR4$, Fig. 8–13) also prevent interaction between any two input circuits that may be in different voltage states. Between any two terminals one of the two diodes in series between the terminals will be biased in the leakage current (high-resistance) region

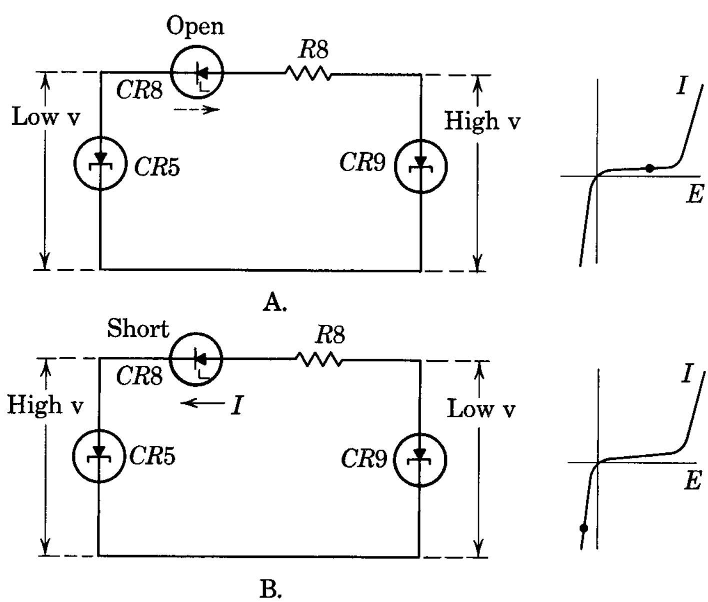


FIG. 8–16. Partial schematic, gate and output diodes, showing effect of different voltage states on backward diode

because the diodes are back-to-back. Similarly backward diodes $CR6$, $CR7$, and $CR8$ prevent interaction between any of the output circuits.

Note: Throughout this text the arrowhead on the graphical symbol for any crystal diode represents p-type semiconductor material. In some technical literature the arrowhead represents p-type material for tunnel diodes and normal rectifying diodes, but then represents n-type material for the backward diode.

8–15. AND or OR Gate, Circuit Analysis

If the basic bistable circuit shown in dashed lines in Fig. 8–13 is biased as indicated in Fig. 8–11B the circuit can be operated as an AND gate (*a* below). If biased as indicated in Fig. 8–11C, the circuit can be operated as an OR gate (*b* below). A discussion of the function of all the parts in the circuit is given in paragraph 8–14.

a. AND Gate. Under quiescent conditions diode $CR5$ (Fig. 8–13) is biased at a low current point in the *on* region. When *all* the preceding tunnel diode stages switch to the *off* state, diode $CR5$ receives sufficient current to exceed its peak current and switches rapidly to the *off* state. In turn the resultant high voltage across diode $CR5$ switches each succeeding stage that is in the *on* state to the *off state.*. To return diode $CR5$ to the *on* state a negative reset pulse is normally introduced at the diode anode. To assure proper operation as an AND gate, several conditions must be met. (Note that in the circuit each input current *adds* to diode $CR5$ current because resistor $R5$ and battery E_B act as a constant-current source that delivers a current I_B at all times. Likewise each output current subtracts from diode $CR5$ current for the same reason.)

1. The sum of the maximum bias current (I_B) plus the maximum values of any two of the input currents (I_1, I_2, or I_3) must be less than the minimum value of the diode $CR5$ peak current (I_p); otherwise the diode will switch with less than *all* inputs present. The terms maximum and minimum are used here, because the indicated current will vary with temperature; in addition they depend on the tolerances of the circuit elements.

2. The sum of the minimum values of all the input currents plus the minimum value of the bias current must be larger than the maximum value of the diode $CR5$ peak current; otherwise the circuit will not switch even when all inputs are present.

3. The minimum value of bias current (I_B) *less* the maximum value of all output currents (I_4, I_5, and I_6) must be greater than the maximum value of the diode $CR5$ valley current (I_v); otherwise the circuit will switch to the *on* state before application of a reset pulse. In practice this condition is usually very difficult to meet with the given circuit. The reason is the

very low value of bias current (I_B) required for AND gate operation. This in turn would require extremely low values of output current. An improved version of this circuit, known as Chow's circuit, is discussed in paragraph 8–16.

b. OR Gate. Under quiescent conditions diode $CR5$ is biased at a high current point in the *on* region. When any one of the preceding tunnel diode stages switch to the *off* state, diode $CR5$ receives sufficient current to exceed its peak current and switches rapidly to the *off* state. In turn, succeeding stages in the *on* state are switched to the *off* state. To assure operation as an OR gate, several conditions must be met:

1. The maximum value of bias current (I_B) must be less than the minimum value of the diode $CR5$ peak current; otherwise the circuit will switch of its own accord to the *off* state.
2. The minimum value of bias current plus the minimum value of any one input current (I_1, I_2, or I_3) must be greater than the maximum value of the diode $CR5$ peak current; otherwise the circuit will not switch to the *off* state with *one* input pulse present.
3. The condition stated in *a*3 above for the AND gate also applies for the OR gate. The difficulty encountered for the practical AND gate, however, normally does not apply to the OR gate because in the latter case bias current (I_B) is much higher.

8–16. Chow's Circuit

The AND gate shown in Fig. 8–17 is named after the originator, W. F. Chow of the General Electric Company. Basically the circuit is similar to that shown in Fig. 8–13 and discussed in paragraphs 8–14 and 8–15. The main difference is the addition of a second tunnel diode ($CR4$) across the

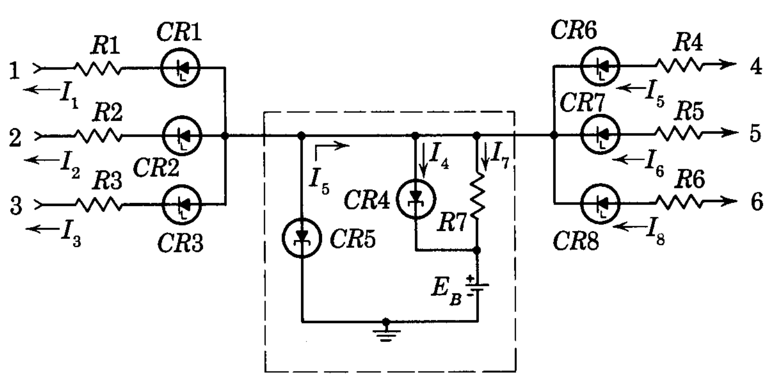


FIG. 8–17. Chow's circuit (Adapted circuit, courtesy General Electric Company)

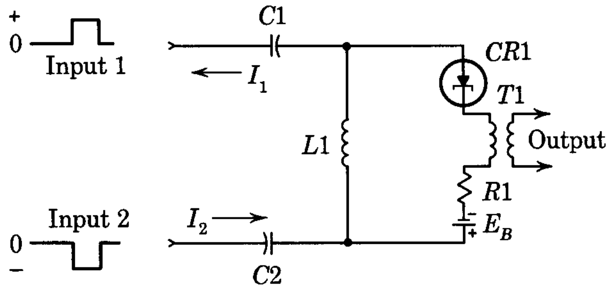
load resistor ($R7$) of the basic bistable circuit shown in dashed lines (Fig. 8–17). This second tunnel diode in turn eliminates the severe output condition experienced with the previous circuit (par. 8–15*a*3). The input and output resistors ($R1$ through $R6$) and the input and output backward diodes perform the same functions as those of the previous circuit.

a. In the previous circuit (Fig. 8–13) the bias current remained a fixed quantity whether diode $CR5$ was in the *on* or the *off* condition. For an AND circuit it is desired to have a low bias current when the diode is in the *on* state and a high bias current when the diode is in the *off* state. With fixed-bias current (constant-current source) only a limited compromise is possible; this condition limits the number of stages that can be fed by the AND gate. Otherwise the possibility of the current through the diode falling below its valley current is risked and unwanted switching occurs.

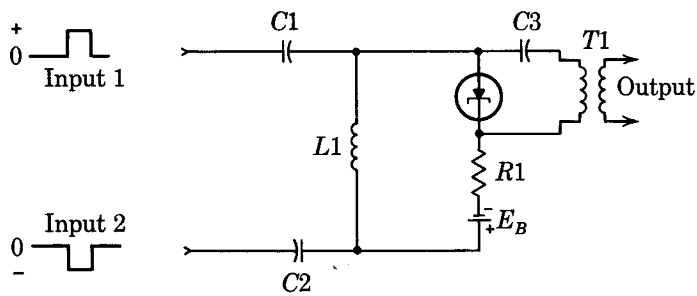
b. Chow's circuit (Fig. 8–17) permits a low bias current to be delivered by battery E_B when diode $CR5$ is in the *on* state, and a high bias current to be delivered by battery E_B when diode $CR5$ is in the *off* state. The voltage value of battery E_B is limited so that only one of the two tunnel diodes ($CR5$ or $CR4$) can be in the high-voltage state; i.e., $E_B = E_p + E_v$ of the diode. In the initial operating condition diode $CR5$ is in the *on* state and diode $CR4$ is in the *off* state. The latter diode draws its low valley current. The total current through diode $CR5$ equals the current through diode $CR4$ and the current through load resistor $R7$. The value of load resistor $R7$ therefore depends upon the number of input pulses. The value is chosen to draw high current for few inputs and very low current for more inputs. When diode $CR5$ is switched to the *off* state by the sum of the input pulses, diode $CR4$ must switch to the *on* state. In the *on* state diode $CR4$ resistance is low and a current almost equal to its peak current can be delivered by battery E_B, plus the current through resistor $R7$. Diode $CR4$ can be selected so that its peak current is much higher than that of diode $CR5$. The current now available for delivery to the following stages (ignoring the current through resistor $R7$) is the difference in diode $CR4$ peak current and diode $CR5$ valley current. To assure AND circuit operation the conditions stated in paragraph 8–15*a*1 and 2 must be met. In addition, the maximum valley current of diode $CR5$ plus the maximum value of all the output currents (I_4, I_5, and I_6) must be less than the minimum peak current of diode $CR4$. Otherwise the circuit will switch to the initial bias condition without application of a reset pulse if diode $CR4$ peak current is exceeded.

8–17. AND Gates, Separate Input Terminals

a. *Input Pulses of Unlike Polarity*. A variation of the basic AND gate is shown in Fig. 8–18A. The basic bistable circuit consists of diode $CR1$,



A. Output proportional to diode current change

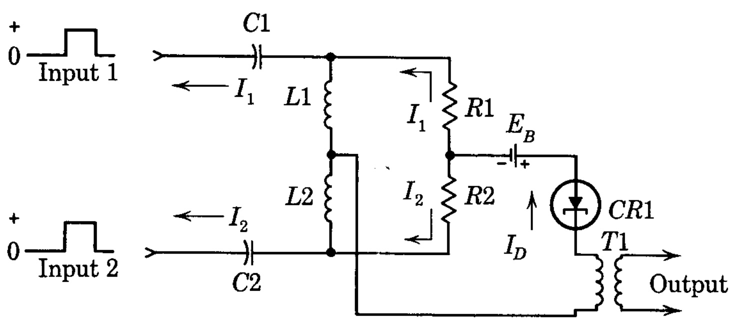


B. Output proportional to diode voltage change

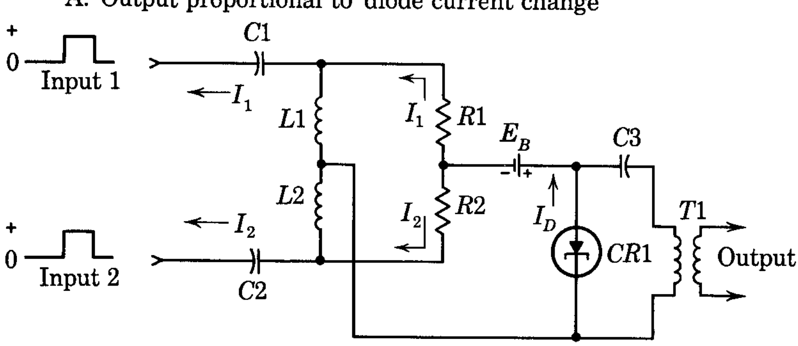
FIG. 8–18. AND gates with separate input terminals and requiring input pulses of unlike polarity

resistor $R1$, and bias battery E_B. The circuit is arranged so that an output will result with one positive input pulse and one negative input pulse. Essentially this condition is obtained by introducing the pulses at opposite ends of the diode. In addition to permitting the use of pulses of opposite polarity, the arrangement offers good isolation of the pulse sources from each other. Capacitors $C1$ and $C2$ are dc blocking capacitors and couple the input pulses to the circuit. Coil $L1$ offers a high ac impedance to the input pulses, and a low dc resistance to the diode bias current. The output is coupled to the following stage through step-up transformer $T1$. Because the diode current flows through the primary of transformer $T1$, the output signal is proportional to the diode current change. Figure 8–18B shows a circuit that is similar in all respects except that the output is proportional to the change in diode voltage. This is accomplished by placing the primary of transformer $T1$ in parallel with the diode through dc blocking capacitor $C3$. AND circuit operation is assured by limiting each input pulse voltage to slightly more than half of the total input voltage required to switch the diode from the *on* to the *off* state.

b. Input Pulses of Like Polarity. Another variation of the basic AND gate is shown in Fig. 8–19A. The basic bistable circuit consists of diode $CR1$, battery E_B, and resistors $R1$ and $R2$; the effective load line is equal to the parallel resistance value of resistors $R1$ and $R2$. Capacitors $C1$ and $C2$ are dc blocking capacitors. Coils $L1$ and $L2$ offer a high ac impedance to the input pulses and a low dc resistance to the diode bias current. The output of step-up transformer $T1$ is proportional to the diode current change. Two positive-going input pulses are required for switching. The pulses are actually fed to the diode anode through isolating parallel networks. Two negative-going input pulses could be used if the battery and diode are connected in opposite directions in the circuit. The circuit shown in Fig. 8–19B is identical in all respects except that step-up transformer $T1$ is in parallel with diode $CR1$ through dc blocking capacitor $C3$. The output therefore is proportional to the voltage change across the diode. AND gate operation is assured by limiting each input pulse current to slightly more



A. Output proportional to diode current change



B. Output proportional to diode voltage change

FIG. 8–19. AND gates with separate input terminals and requiring input pulses of like polarity

than half the total input current required to switch the diode from the *on* to the *off* state.

8–18. NOT Gate or EXCLUSIVE OR Gate

The circuit shown in Fig. 8–20 can be used as a NOT gate or an EXCLUSIVE OR gate. As a NOT gate (*a* below) it is required that the peak current of diode $CR1$ be relatively low. As an EXCLUSIVE OR gate (*b* below), it is required that the peak current of diode $CR1$ be relatively high. Either application is best analyzed by graphical means. Resistor $R1$ is the pulse-source isolating resistor if the circuit is used as a NOT gate. Resistors $R2$ and $R3$ are pulse-source isolating resistors if the circuit is used as an EXCLUSIVE OR gate. Bias battery E_B and resistor $R4$ form a constant-current source that delivers a fixed current (I_B). Diode $CR1$ is considered the diode that is being switched, whereas diode $CR2$ and resistor R_L, in series, are considered to form a *load* for diode $CR1$. The output is taken across resistor R_L. The effective *load* formed by diode $CR2$ and resistor R_L is obtained by considering the graphs shown in Fig. 8–21. Refer to Fig. 8–20 for the current and voltage designations used. In Fig. 8–21A the current through and the voltage across diode $CR2$ (I_2 vs E_2) is

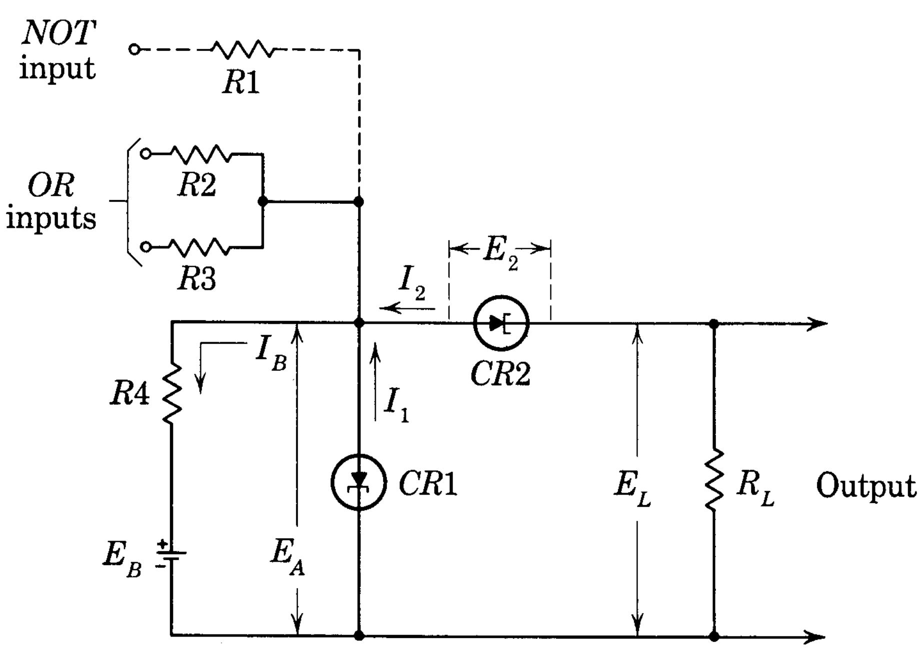


Fig. 8–20. NOT gate or EXCLUSIVE OR gate

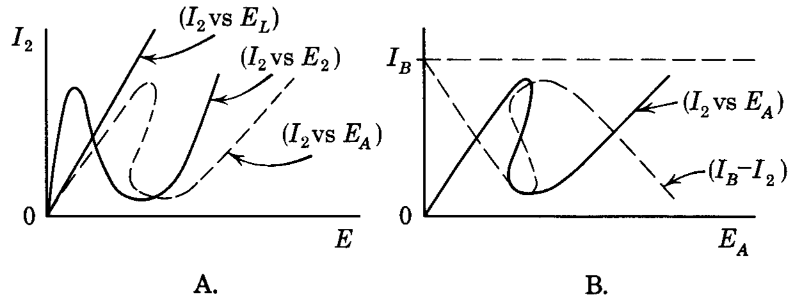


FIG. 8–21. Development of an effective parallel load line formed by a tunnel diode and a resistor

drawn; the current through and the voltage across resistor R_L (I_2 vs E_L) is drawn. Note that the current through diode $CR2$ and resistor R_L is the same and that the applied voltage (E_A) equals the sum of the voltages across these two elements ($E_A = E_2 + E_L$). If diode and resistor curves are added point for point in a horizontal direction, their composite curve (I_2 vs E_A) is obtained. In Fig. 8–21B the composite curve (I_2 vs E_A) is subtracted from the fixed bias current (I_B) and the difference ($I_B - I_2$) represents the effective load. Actually $I_B - I_2$ equals I_1. When the curve ($I_B - I_2$) is superimposed on diode $CR1$ current-voltage curve, the points of intersection in the positive-resistance regions will be the stable operating points; i.e., those points are the only points that represent a *compatible* division of bias current between the two branches of the circuit.

a. NOT Gate. Figure 8–22A shows diode $CR1$ curve (I_1) and the effective load ($I_B - I_2$) formed by diode $CR2$ and resistor R_L. Note that diode $CR1$ has a low peak current. Points 1 and 2, both in the high-voltage region of diode $CR1$, are the only stable points. When the power is first turned on, the circuit stabilizes at point 1. Here the current through diode $CR1$ is low;

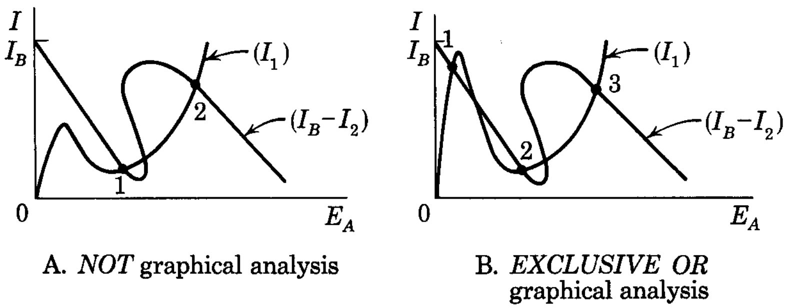


FIG. 8–22. Biasing relationships for NOT gate and EXCLUSIVE OR gate

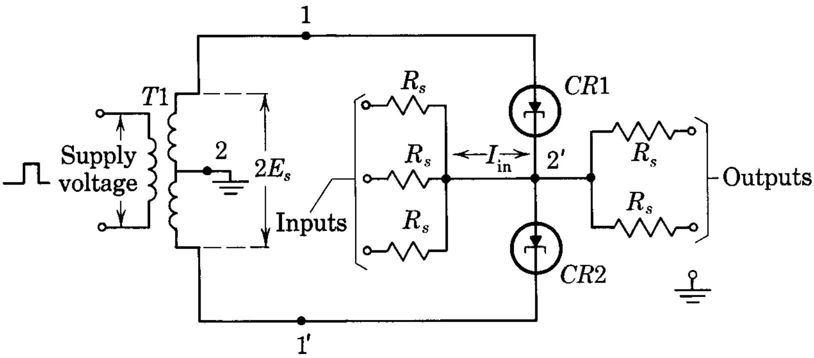
that through resistor R_L is high and the output voltage is a high positive value. A positive input current switches the circuit to point 2. Here the current through diode $CR1$ is high; that through resistor R_L is low and the output voltage is low. A negative input current now has the opposite effect. The circuit therefore performs the NOT or inversion function.

b. EXCLUSIVE OR Gate. Figure 8–22B shows diode $CR1$ curve and the effective load $(I_B - I_2)$. Note that in this case diode $CR1$ has a high peak current which results in *three* stable operating points. At point 1 the current through resistor R_L is low and the output voltage is low. If *either* input pulse of high magnitude is present, the circuit will switch to point 2 which is a high-voltage output point. If *both* input pulses of high magnitude are present, the circuit will switch to point 3 which is a low-voltage output point. The EXCLUSIVE feature refers to whether the output will be a high- or a low-voltage output.

8–19. MAJORITY Gate, Goto Pair

The MAJORITY gate shown in Fig. 8–23A is named after the originator, E. Goto of the University of Tokyo. Basically the circuit consists of the coupled-pair or twin-diode arrangement, the current-voltage characteristic of which is discussed in detail in paragraph 6–7. The supply voltage $(2E_s)$, however, consists of a pulse coupled into the circuit through transformer $T1$. During operation the supply voltage pulse is rising when the trigger input pulses are introduced. The input pulses are coupled through isolating resistors R_s. The outputs are taken through isolating resistors R_s. These resistors (considered to be all in parallel and of equal value) also form a bistable load line with the current-voltage characteristic of terminals 2–2′ (Fig. 8–23B). The circuit is stable at points 1 and 2. Note that an odd number of inputs are used. Both negative and positive input trigger pulses are used. If the sum of the input currents causes a net electron flow (I_{in}) into terminal 2′, the circuit will stabilize at point 2 and the output voltage will be positive with respect to ground. If the sum of the input currents causes a net electron flow (I_{in}) out of terminal 2′, the circuit will stabilize at point 1 and the output voltage will be negative. In other words, the polarity of the output voltage depends upon the polarity of the majority of the input pulses. The circuit is reset each time the supply voltage pulse drops to zero.

Note: The Goto pair is actually one example of tunnel diode *locking* circuits. Locking circuits usually do not have a dc power supply and are receptive to input signals only during the time that the power supply waveform rises from zero to its maximum value. The circuit then *locks* into one of its two possible stable states. Another example of a locked pair is discussed in paragraph 8–20.



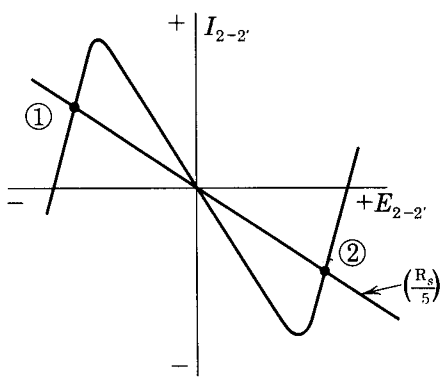


FIG. 8–23. Goto pair MAJORITY gate, and bistable arrangement of characteristics

8–20. Majority Gate, Single-Ended Locked Pair

a. Like the Goto pair (par. 8–19), the circuit shown in Fig. 8–24 is classified as a locking circuit; the input power is in the form of a rectangular pulse coupled into the circuit by transformer $T1$. Unlike the Goto pair, the output is taken from anode to cathode of one tunnel diode and is therefore a *single-ended* circuit. The secondary of the input transformer need not be center-tapped. In this circuit, the two diodes are matched. Resistors $R1$, $R2$, and $R3$ are the input isolating resistors. Supply voltage E_s is limited so that only one of the two diodes can be in the *off* state. The circuit therefore is stable with diode $CR1$ in the *off* state and diode $CR2$ in the *on* state, or vice versa.

b. Pulses of mixed polarity are introduced into terminals 1, 2, and 3 at

the same time that the power supply pulse is rising in magnitude. Current I_1 in diode $CR1$ and I_2 in diode $CR2$ start to rise.

1. If negative input pulses predominate, input current I_{in} flows *toward* the junction of the diodes; this current adds to current I_1 and causes diode $CR1$ to exceed is peak current and switch rapidly to its *off* state. Diode $CR1$ remains in the *on* state and the output voltage is low.
2. If positive input pulses predominate, input current I_{in} flows *away* from the junction of the diodes. This current adds to current I_2 and causes diode $CR2$ to exceed its peak current and switch rapidly to its *off* state. The output voltage is high.
3. Thus the magnitude of the output voltage depends upon the polarity of the *majority* of the input pulses.

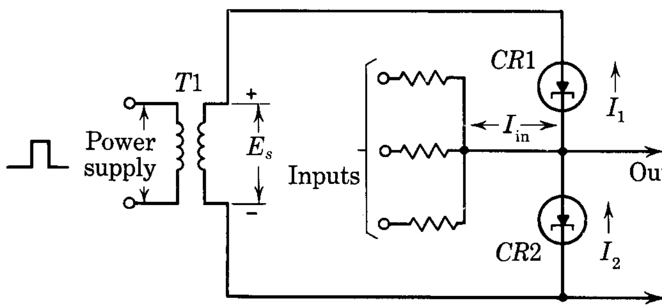


FIG. 8–24. Single-ended, locked pair MAJORITY gate

8–21. AND or OR Locking Circuit

The locking circuit shown in Fig. 8–24 can also be designed for use as an AND gate or an OR gate. This can be accomplished by using two tunnel diodes having different peak currents. As an AND gate (*a* below) the difference in peak currents must be relatively large. As an OR gate (*b* below) the difference in peak currents must be relatively small. The power supply limitations are the same as for the MAJORITY gate (par. 8–20).

a. AND Gate. Assume that the diode $CR1$ has a much lower peak current than diode $CR2$. With no input pulses except from the power supply, diode $CR1$ will always reach its peak current first and switch to the *off* state. Diode $CR2$ must remain in the *on* state and the output voltage is low. If all the input pulses of predetermined value are present and positive-going (drawing current from the junction of the diodes) when power is turned on, the aiding effect on diode $CR2$ will cause it to switch to the *off* state first. The output voltage is high. AND circuit operation is assured by having the difference in the diode peak currents sufficiently large that *all* inputs must be present to cause diode $CR2$ to switch first. All negative-going input

pulses can also be used. In this case diode $CR1$ will have a *higher* peak current than diode $CR2$. When all input pulses are present, the output voltage will be low instead of high as with positive-going input pulses.

b. OR Gate. The operation of the locking circuit as an OR gate is identical with its operation as an AND gate (*a* above). The peak currents of the diodes, however, are only slightly different in magnitude, and effective switching can be accomplished with only one input pulse present.

8–22. Summary

a. Switching circuits perform triggering, gating, inverting, and signal routing functions.

b. Pulse and switching circuits are characterized by large-signal operation.

c. A unit-step voltage (or current) refers to an instantaneous change in amplitude either positive- or negative-going.

d. When biased in the low-voltage state, the tunnel diode is considered to be *on;* in the high-voltage state, it is *off*.

e. The basic monostable multivibrator requires a bias voltage, a load resistor, a tunnel diode, and an energy-storing element, such as a coil. The monostable multivibrator may be stable in the *on* region or the *off* region.

f. The basic bistable multivibrator requires a bias voltage, a load resistor, and a tunnel diode. The bistable multivibrator is stable in the *on* region and the *off* region.

g. The basic bistable multivibrator can be made more sensitive by shunting the tunnel diode with a normal rectifying diode in series with a low-valued resistor (par. 8–7). The resultant output waveform has almost equal dwell periods. The base-emitter junction of a transistor can be substituted for the rectifying diode.

h. Monostable or bistable multivibrators can be constructed from a coupled pair. Trigger pulses of the *same* polarity can be used for switching in either direction (par. 8–8).

i. A controlled-negative-resistance multivibrator (par. 8–9) produces a variable-amplitude output waveform.

j. By using a multistate switch composed of several tunnel diodes in series, a pulse-frequency divider and staircase generator can be formed (par. 8–10).

k. Most AND or OR gates are similar except for the bias points used in the basic bistable circuit.

l. Inversion of the output of an AND or OR gate results in a NOT AND or NOR gate, respectively.

m. Unidirectional flow of energy (or information) is achieved in certain gating circuits (par. 8–14) by using backward diodes at the input and output terminals.

n. A backward diode may be considered a *low-voltage* rectifying diode.

o. A NOT gate can be formed by using a tunnel diode and a series resistor combination as a parallel load for the *active* tunnel diode (par. 8–18).

p. The EXCLUSIVE OR gate uses similar components as the NOT gate (*o* above). The NOT gate has two stable operating points; the EXCLUSIVE OR gate has *three* stable operating points.

q. A locking circuit is one that is receptive to input pulses only during the rising portion of the power supply waveform.

r. The Goto pair MAJORITY gate is a special application of the coupled pair in a locking circuit.

s. By using two tunnel diodes in a single-ended locking circuit (par. 8–20 and 8–21), a MAJORITY gate, and AND gate, or an OR gate can be formed.

Chapter 9

MODULATORS, DEMODULATORS, AND HETERODYNE DETECTORS

SECTION I. MODULATORS

9–1. Modulation, General

The process of varying a particular characteristic of radio signal (a carrier) in accordance with the amplitude of a signal that represents intelligence, such as speech or music, is called *modulation*. The most commonly used types of modulation are called amplitude modulation (AM) and frequency modulation (FM); these processes are discussed briefly in paragraphs 9–2 and 9–5, respectively.

9–2. Amplitude Modulation

The basic process of amplitude modulation (AM) is represented in block form in Fig. 9–1. The amplitude of a *carrier* (RF signal) is varied in accordance with the amplitude of a *modulating signal*. The latter signal, usually of low frequency (20 cps to 20 kc), is an electrical representation of the intelligence to be transmitted. The most direct method of producing an amplitude-modulated carrier uses the technique of varying the gain of

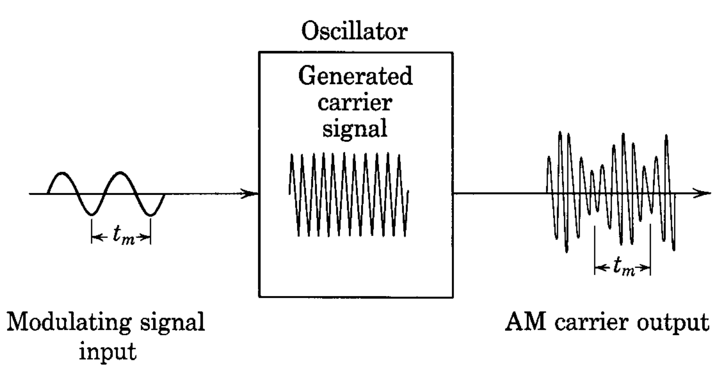


FIG. 9–1. Block diagram of amplitude-modulated oscillator

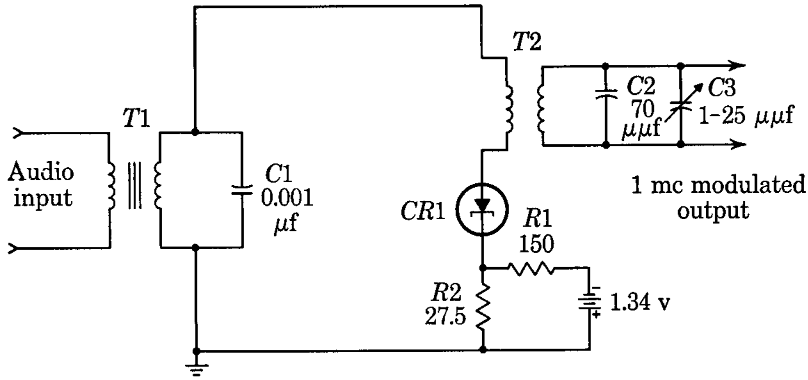


FIG. 9–2. Amplitude-modulated, 1-mc tunnel diode oscillator

the oscillator (that generates the carrier signal) in accordance with the amplitude of the modulating signal. Because the negative-resistance characteristic of the tunnel diode facilitates the construction of an oscillator, this technique is most often employed. Typical tunnel diode, amplitude-modulated oscillators are covered in paragraphs 9–3 and 9–4.

9–3. Single Diode AM Oscillator

A single diode amplitude-modulated oscillator is shown in Fig. 9–2. The frequency of oscillation (1 mc) is determined mainly by the tank circuit consisting of the transformer $T2$ secondary in resonance with capacitor $C2$. Trimmer capacitor $C3$ permits tuning of the tank circuit to the desired frequency. Diode $CR1$ bias is supplied through the voltage divider consisting of resistors $R1$ and $R2$. The bias for diode $CR1$ is varied at an audio rate by the signal introduced into the circuit by transformer $T1$.

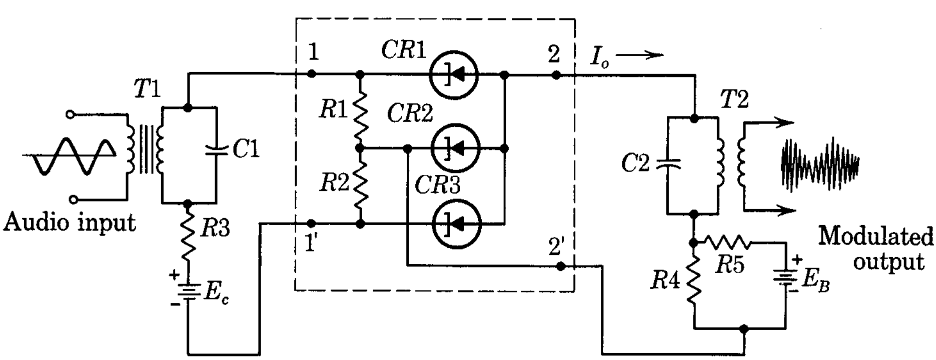


FIG. 9–3. Amplitude-modulated oscillator using three tunnel diode controlled-negative-resistance *device*

Capacitor $C1$ bypasses the 1-mc carrier signal around the audio circuit. The amplitude of oscillation varies in accordance with the input modulating signal. The resultant output is a 1-mc, amplitude-modulated carrier signal.

9–4. Amplitude-Modulated Controlled-Negative-Resistance Device

An amplitude-modulated oscillator is shown in Fig. 9–3. In this circuit the combination of diodes $CR1$, $CR2$, and $CR3$, and resistors $R1$ and $R2$ (shown in dashed lines) form a controlled-negative-resistance *device*. Typical characteristics of the *device* were considered in paragraph 6–9. The family of characteristic curves for the arrangement used in the amplitude-modulated oscillator is shown in Fig. 9–4. Note that the negative resistance value at terminals 2–2′ depends upon the control current (I_c) introduced at terminals 1–1′. In the circuit shown, resistor $R3$ has a high value and, in conjunction with battery E_c, simulates a constant-current source. At quiescence it introduces a control current of 0.75 ma at terminals 1–1′; the characteristic curve at this value is shown. Battery E_B in conjunction with the voltage divider (resistors $R4$ and $R5$) biases the output circuit approximately at the midpoint of the negative-resistance

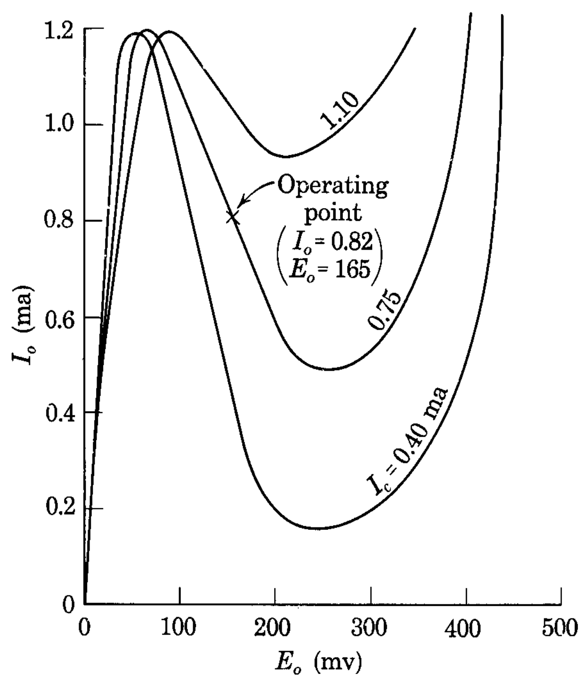


FIG. 9–4. Bias point indicated on three tunnel diode characteristic curves

portion of the 0.75-ma curve. With the positive resistance (in the output circuit) less than the total negative resistance, the output circuit will oscillate at an RF frequency determined mainly by the tank circuit consisting of capacitor $C2$ and the primary of transformer $T2$. If an audio signal is introduced into the input circuit through transformer $T1$ as shown, the control current will be varied at an audio rate. Increased control current (to 1.10 ma) forces the oscillator to operate on the top curve (Fig. 9–4), with a minimum output from the oscillator. Decreased control current (to 0.40 ma) forces the oscillator to operate on the lower curve, with a maximum output from the oscillator. The result is an audio modulated RF output. In the circuit capacitor $C1$ is an RF bypass capacitor which prevents feedback of RF to the audio circuit. Note that the two tunnel diode controlled-negative-resistance *device* (par. 6–8) can also be used in a similar circuit.

9–5. Frequency Modulation

The basic process of frequency modulation is represented in block form in Fig. 9–5. The frequency of a carrier (RF signal) is varied in accordance with the amplitude of a modulating signal; the rate of frequency change of the carrier is proportional to the frequency of the modulating signal. The latter signal is an electrical representation of the intelligence to be transmitted. Frequency modulation is established by varying the frequency of the oscillator (that generates the carrier signal) in accordance with the amplitude of the modulating signal and at a rate determined by the frequency of the modulating signal. The amplitude of the transmitted signal remains constant. In the illustration the negative peaks of the modulating signal are shown separated by time t_m. The relative position of this time factor in the FM signal is also marked by t_m. Note that the modulating signal negative peaks result in a low frequency in the FM signal; the posi-

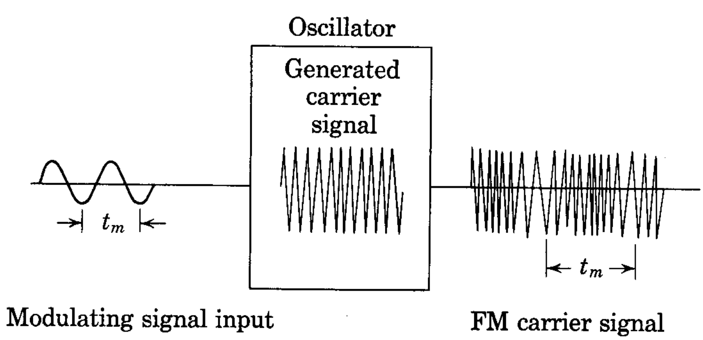


FIG. 9–5. Block diagram of frequency-modulated oscillator

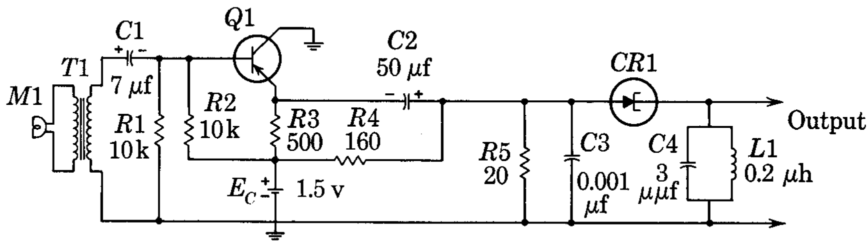


FIG. 9–6. Frequency-modulated (100-mc) oscillator (Adapted circuit, courtesy General Electric Co.)

tive peaks result in a high frequency. A typical tunnel diode FM oscillator is discussed in paragraph 9–6.

9–6. FM Oscillator

a. A practical 100-mc, frequency-modulated tunnel diode oscillator is shown in Fig. 9–6. The circuit consists of two stages, transistor *Q*1 audio amplifier and the diode oscillator (*CR*1). The audio amplifier is an emitter follower that amplifies the weak output from dynamic microphone *M*1. Transformer *T*1 couples the output of microphone *M*1 to the base-emitter circuit of transistor *Q*1. Capacitor *C*1 is a blocking capacitor that prevents shorting of the base-bias voltage to ground through the secondary of transformer *T*1. Resistors *R*1 and *R*2 form a voltage divider that provides the forward-bias voltage. Resistor *R*3 is the emitter load resistor that develops the amplified audio current output. Battery E_C provides bias voltage for the transistor and the diode. Dc blocking capacitor *C*2 couples the audio signal to the diode anode. Resistors *R*4 and *R*5 form a low-impedance voltage divider to establish the diode bias. The frequency of the diode oscillator is determined mainly by the tank circuit consisting of capacitor *C*4 and coil *L*1. RF bypass capacitor *C*3 prevents feedback from the 100-mc oscillator stage to the audio stage.

b. It was shown in paragraph 3–17 that the frequency of oscillation of a tunnel diode oscillator depends upon the negative-resistance value of the tunnel diode. Because the negative-resistance value varies slightly with bias (par. 4–6), the frequency of oscillation varies with bias. In this case the bias is being varied at an audio rate. The resultant output is a frequency-modulated signal: the positive audio peak *increases* the oscillator frequency by 75 kc; the negative audio peak *decreases* the oscillator peak by 75 kc. The frequency deviation therefore is ± 75 kc. This circuit can be used as a wireless microphone and can operate efficiently within 100 ft of an FM receiver having a minimum sensitivity of 10 μv.

SECTION II. DEMODULATORS

9–7. Demodulation

The basic process of demodulation is illustrated in Fig. 9–7. An amplitude-modulated carrier is introduced into an AM detector (demodulator). The output is the original modulating signal. Demodulation then is the process of extracting the original modulating signal from a modulated carrier.

a. In the case of electron tubes and transistors demodulators are classified as *linear* detectors and *square-law* detectors. The linear detector produces an output magnitude which is directly proportional to the input magnitude.

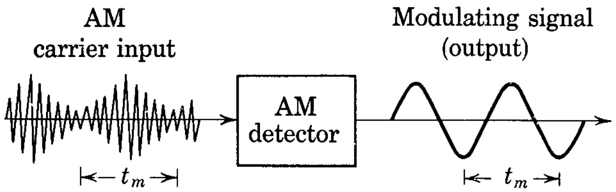


FIG. 9–7. Block diagram, AM detector, showing input and output waveforms

A square-law detector produces an output magnitude which is proportional to the *square* of the input magnitude. The tunnel diode, used as an AM demodulator, displays the characteristics of a square-law detector. A brief discussion of an electron-tube square-law detector is given in paragraph 9–8. The tunnel diode AM detector is discussed in paragraph 9–9.

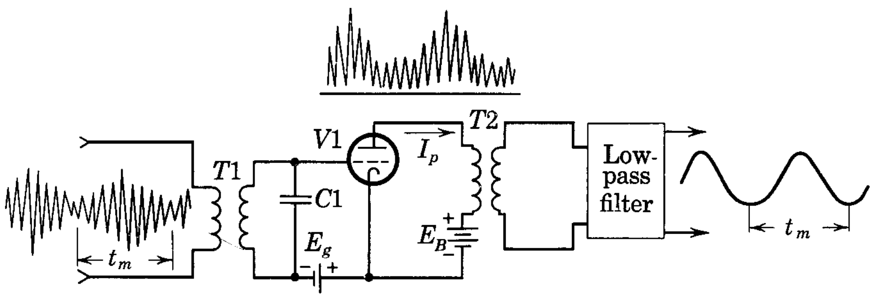
b. To extract the original modulating signal from an FM carrier, an FM detector is required. The tunnel diode can be used to demodulate FM carrier signals by use of a technique referred to as slope detection. A circuit employing this technique is discussed in paragraph 9–10.

9–8. Electron-Tube Square-Law Detector

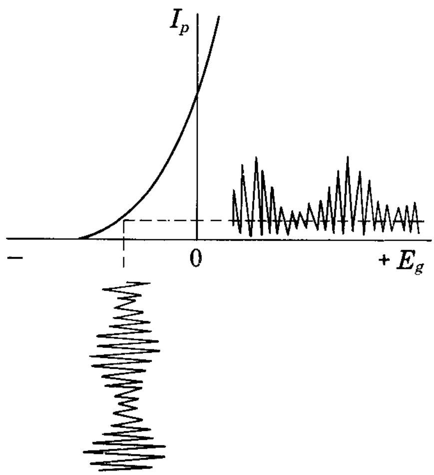
The operation of an electron-tube square-law detector (Fig. 9–8A) is discussed here so that the operation of the tunnel diode detector (par. 9–9) can readily be compared.

a. Although the amplitude of a carrier signal may vary in accordance with the amplitude of a modulating signal, the AM carrier does *not* contain the original modulating signal. The AM carrier contains the original carrier and upper and lower sidebands removed from the carrier frequency by the frequency of the modulating signal. The original modulating signal can be *created* by deliberately distorting the AM carrier by full- or half-wave rectification, or by nonlinear amplification of the carrier signal. The

square-law detector operates on the latter principle. The electron tube is biased so that operation occurs on the curved portion of the plate-current (I_p)-grid voltage (E_g) curve (Fig. 9–8B). Note that one half of the AM carrier receives greater amplification than the other half. The resultant



A. Electron-tube square-law detector



B. Biasing

FIG. 9–8. Electron-tube square-law detector, showing method of biasing

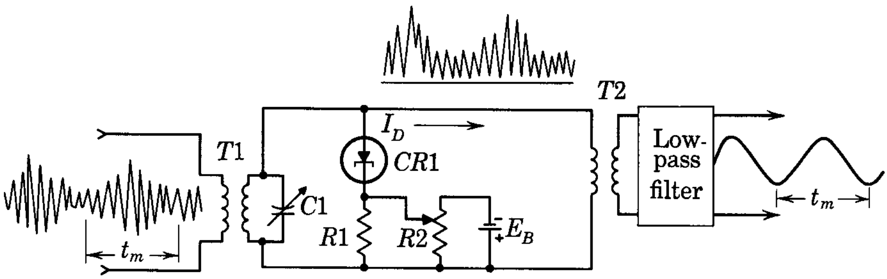
plate current now contains the original modulating signal. If the plate current is passed through a low-pass filter, the modulating signal is recovered. Because a portion of the I_p–E_g curve used approximates a parabola, the output is proportional to the square of the input.

b. In the circuit shown (Fig. 9–8A), transformer $T1$ couples the AM carrier into the grid-cathode circuit. Battery E_g biases the grid-cathode circuit. Capacitor $C1$ resonates with the transformer $T1$ secondary at the

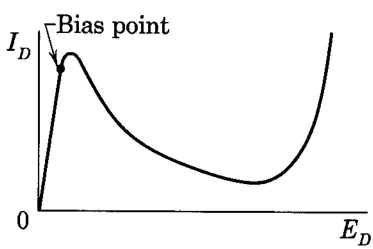
carrier frequency. The distorted carrier plate current is indicated. Transformer $T2$ couples the output of tube $V1$ to a low-pass filter which recovers the modulating signal. Battery E_B biases the cathode-plate circuit.

9–9. Tunnel Diode Square-Law Detector

A tunnel diode square-law detector is shown in Fig. 9–9A. Transformer $T1$ couples the AM carrier to the diode. Capacitor $C1$ tunes the transformer



A. Tunnel diode square-law detector



B. Biasing

FIG. 9–9. Tunnel-diode square-law detector and point of bias

$T1$ secondary to the carrier signal. Resistors $R1$ and $R2$ form a voltage divider to bias diode $CR1$. Transformer $T1$ couples the output of diode $CR1$ to a low-pass filter which extracts the original modulating signal. As in the case of the electron-tube square-law detector (par. 9–8), the tunnel diode is biased so that the upper portion of the AM carrier receives higher amplification than the lower portion. (Actually, the lower portion receives no amplification.) The point of bias of the tunnel diode is just below the peak-current point (Fig. 9–9B). Note that the output current (I_D) of diode $CR1$ (Fig. 9–9A) is similar in form to the output current (I_p) of the electron tube (Fig. 9–8A).

a. For an explanation of the operation of the tunnel diode, refer to Fig. 9–10. For clarity the *idealized* current-voltage characteristic of the tunnel

diode is represented by curve A. The effective load on the tunnel diode is represented by curve B. The composite of curves A and B is represented by curve C. Note that these curves are identical to those used for graphical analysis of the parallel amplifier (par. 3–24). The bias point is represented by a dot on curves A and C.

b. To determine the resultant amplification, the AM carrier input current is projected onto composite curve A. To determine the current in the effec-

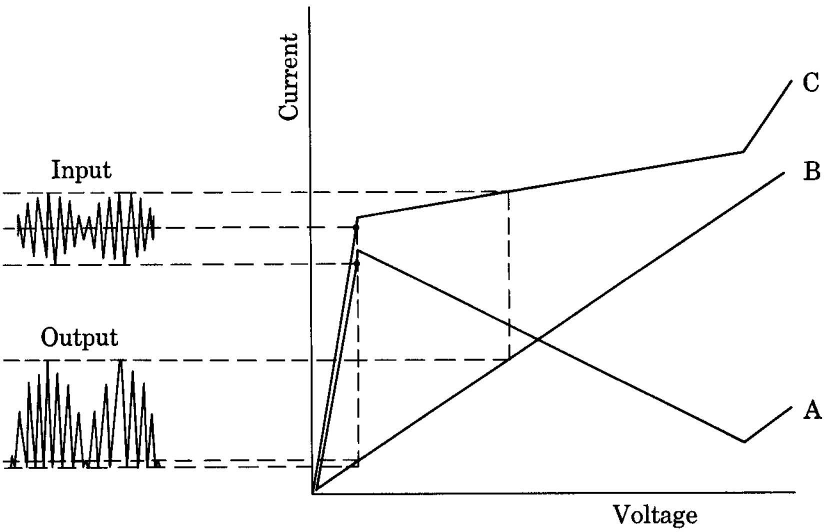


FIG. 9–10. Graphical analysis used to show detection by properly biased tunnel diode

tive load, the points of intersection on composite curve A are projected vertically down until they intersect the effective load line (curve B). These points in turn are projected horizontally to the left to obtain the output current. Note that the upper portion of the AM carrier signal has experienced large current gain because this portion extends into the negative-resistance region of the tunnel diode. If the practical current-voltage curve of the tunnel diode were used instead of the idealized curve, it can be shown that the output is proportional to the square of the input; thus permitting square-law detection.

9–10. FM Slope Detector

A possible circuit arrangement for detecting FM carrier signals is shown in Fig. 9–11A. Essentially the circuit converts the FM carrier to an AM carrier. The AM carrier is then demodulated by the square-law tunnel diode detector. The functions of transformers $T1$ and $T2$, resistors $R1$ and

$R2$, and diode $CR1$ are identical with the correspondingly referenced circuit elements shown in Fig. 9–9A. However, the parallel resonant circuit consisting of coil $L1$ and capacitor $C1$ has been added in series with the secondary of transformer $T1$. The curve of impedance versus frequency for this resonant circuit is shown in Fig. 9–11B. Note that the peak (resonant frequency point) of the curve is well above the FM carrier frequency. The FM carrier frequency falls approximately at the center of the lower *slope*. The result is a higher impedance for the high-frequency deviations of the carrier, and a lower impedance for the low-frequency deviations of the carrier. Correspondingly, the voltage input to the diode is greater for the higher frequencies (originally caused by the positive half of the modulating signal) than for the lower frequencies (originally caused by the negative half of the modulating signal). The result is a conversion of the FM carrier to an AM carrier. The circuit then functions as an AM detector

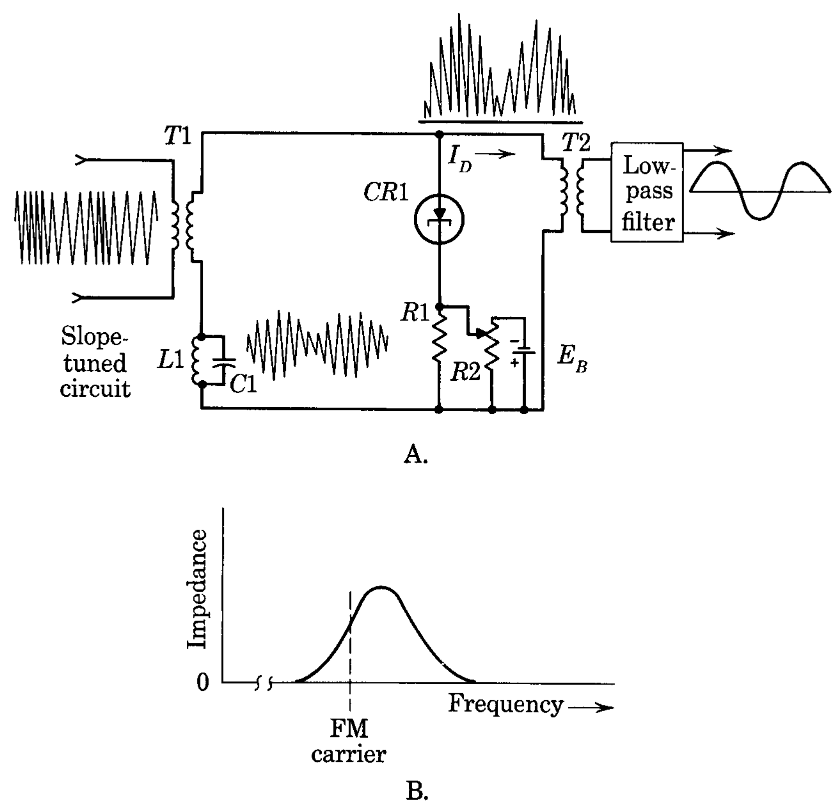


FIG. 9–11. FM slope detector showing position of FM carrier on impedance-frequency curve of parallel tuned circuit

(par. 9–9). Because of the position of the FM carrier on the slope of the impedance-frequency curve, the overall circuit is referred to as a *slope* detector.

SECTION III. HETERODYNE DETECTORS

9–11. Heterodyne Detection

The basic principle of heterodyne action is demonstrated in Fig. 9–12. If two signals (E_s and E_o) of slightly different frequencies are added or superimposed in a given circuit, the result will be a signal (E_r), the amplitude of which will vary at a rate equal to the difference in frequency between E_s and E_o. The amplitude modulation of the resultant wave (E_r) at the difference frequency is occasioned by the periodic in-phase and out-of-phase relationship between E_s and E_o. The percentage of modulation of the resultant wave depends upon the relative magnitude of E_s and E_o; if equal, 100% modulation occurs; if one is one half of the other, 50% modulation results.

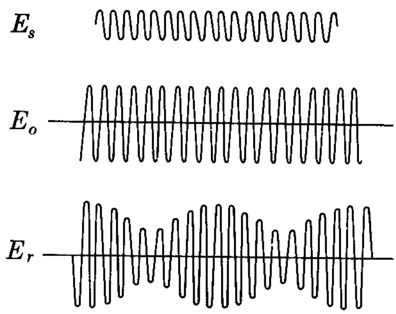


FIG. 9–12. Combination of two signals (E_s and E_o) to produce a resultant signal (E_r), amplitude modulated at the difference frequency

a. The resultant signal (E_r) contains only *two* frequencies, the original frequencies (E_s and E_o). To produce the *difference frequency,* signal E_r must be deliberately distorted by rectification or square-law detection (par. 9–8 and 9–9). Such distortion produces a signal that contains the difference frequency, the original frequencies (E_s and E_o), as well as the sum frequency ($E_s + E_o$); other frequencies may also be present.

b. Production of the difference frequency or the sum frequency is called heterodyne detection. The most common application of this process is in superheterodyne receivers. In this case the difference frequency is of most interest. The received radio signal (a high frequency for ease of transmission) is heterodyned against a signal generated in the receiver by the local oscillator to produce a lower frequency [intermediate frequency (IF) for ease of amplification].

c. If the local oscillator signal is generated by a separate circuit and coupled into the heterodyne detector along with the received radio signal, the heterodyne detector is referred to as a *mixer.* If the heterodyne detector also generates the local oscillator signal, the heterodyne detector is referred to as a *converter.*

d. A mixer circuit using a normal rectifying diode is discussed in paragraph 9–12. Mixer circuits using tunnel diodes are discussed in paragraphs 9–13 and 9–14. Tunnel diode converters are discussed in paragraph 9–15.

9–12. Mixer, Using Normal Diode Rectifier

A very simple mixer stage is shown in Fig. 9–13. The RF signal (E_s) and the local oscillator signal (E_o) are combined in series by the series

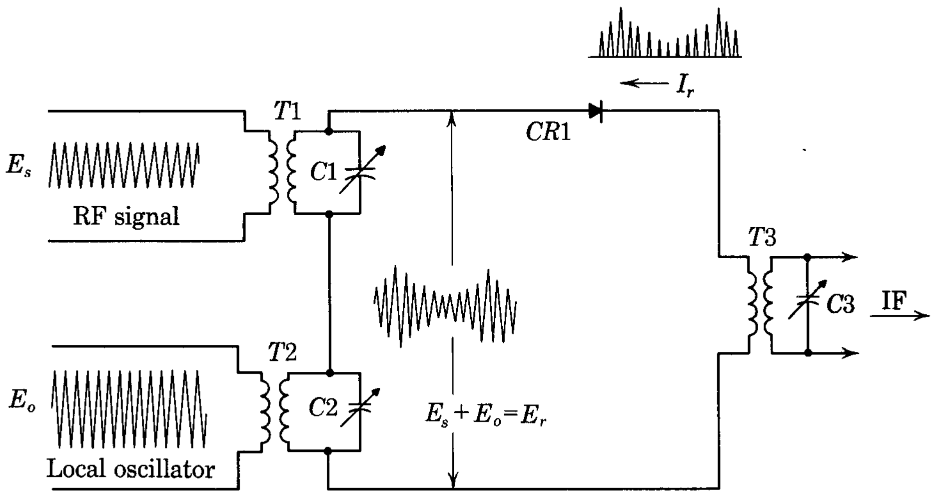


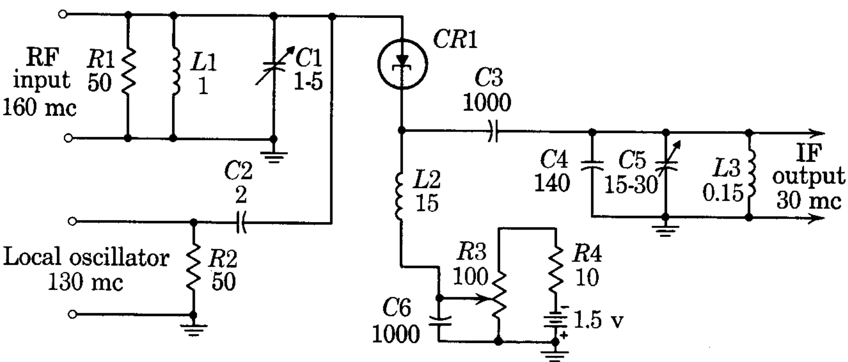
FIG. 9–13. Normal rectifying diode used in a mixer stage

arrangement of the secondaries of transformers $T1$ and $T2$ to produce a heterodyne signal (E_r). Capacitors $C1$ and $C2$ form tuned circuits with the secondaries of transformers $T1$ and $T2$ at the RF signal and the local oscillator signal, respectively. The normal-diode ($CR1$) rectified current (I_r) contains a number of signals (par. 9–11) including the IF frequency. With transformer $T3$ secondary tuned to this frequency, only the IF frequency is coupled to the following stage. One disadvantage of the normal rectifier mixer circuit is that no conversion gain results. The tunnel diode mixer (par. 9–13) does produce gain.

9–13. Tunnel Diode Mixer

A tunnel diode mixer stage is shown in Fig. 9–14. A 160-mc RF input and a 130-mc local oscillator input are heterodyned in diode $CR1$. The output is the 30-mc IF frequency. To operate as a heterodyne detector, the diode is biased just below the peak current as indicated by the dot on the diode current-voltage characteristic shown in Fig. 9–9B. Except for the

introduction of the local oscillator signal to produce heterodyne action (par. 9–11), the operation of the tunnel diode mixer is similar to the operation of the tunnel diode (square-law detector) demodulator. (See par. 9–9 and Fig. 9–10.) Because the mixer operates over a portion of the negative-resistance region of the diode characteristic, conversion gain (with relatively low noise) results. The circuit shown produces a conversion gain of 3 to



Note: Resistances in ohms
Capacitances in $\mu\mu$f
Inductances in μh

FIG. 9–14. Tunnel diode mixer stage

6 db, depending upon the exact point of bias. The circuit can be divided into four parts, RF input, local oscillator input, bias circuit, and IF output; these are discussed below.

a. Resistor *R*1 matches the output impedance of the RF input source to the tunnel diode. Capacitor *C*1 and coil *L*1 form a parallel resonant circuit that can be tuned to 160 mc.

b. Resistor *R*2 is an impedance-matching resistor. Capacitor *C*2 is a dc blocking capacitor that couples the local oscillator signal to diode *CR*1.

c. Resistors *R*3 and *R*4 form a voltage divider to bias diode *CR*1. Bypass capacitor *C*2 and choke coil *L*2 decouple the bias source from the diode *CR*1 cathode.

d. Dc blocking capacitor *C*3 couples the ac output of diode *CR*1 to the IF resonant tuned circuit consisting of capacitor *C*4 and coil *L*3; trimmer capacitor *C*5 permits tuning exactly at 30 mc. This resonant circuit also acts as a short circuit to all other frequency components of the diode output.

9–14. Hybrid-Coupled Tunnel Diode Mixer

a. In many applications a tunnel diode mixer is fed from an RF tunnel diode amplifier as well as a tunnel diode local oscillator. In these applications it has been found that even if a small amount of the oscillator output reaches the RF amplifier, the gain of the amplifier is substantially reduced. To avoid this condition, it is necessary to isolate the RF amplifier from the output of the local oscillator. A circuit arrangement which successfully achieves this requirement is shown in Fig. 9–15. Isolation of the RF amplifier from the local oscillator is obtained by use of a hybrid junction. A brief

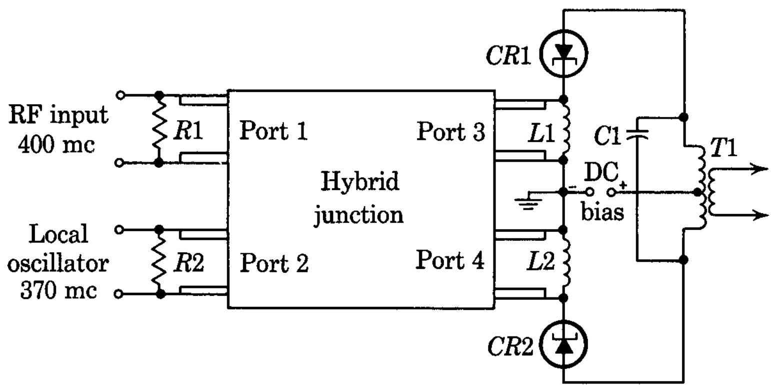


FIG. 9–15. Hybrid-coupled tunnel diode mixer

discussion of the theory of operation of the hybrid junction is presented in paragraph 5–18. Essentially the hybrid junction is a four-port waveguide device that prevents exchange of energy between adjacent ports (i.e., ports 1 and 2, or ports 3 and 4). By placing the RF amplifier and the local oscillator at adjacent ports as shown, the amplifier is isolated from the oscillator. Resistors *R*1 and *R*2 terminate ports 1 and 2, respectively, in their characteristic impedance for maximum effective hybrid action.

b. At ports 3 and 4 two matched tunnel diodes (*CR*1 and *CR*2) operating in push-pull through center-tapped output transformer *T*1 operate as mixers. The tunnel diodes are biased near the peak point of their characteristic (Fig. 9–9B) and operate in the same manner as the single-tunnel diode mixer (par. 9–13). The mixer stage (Fig. 9–15) heterodynes the 400-mc RF signal and the 370-mc local oscillator signal. The resultant signal is distorted and detected by the diodes to produce the IF (30-mc) signal. Capacitor *C*1 tunes the transformer *T*1 primary to 30 mc. The diodes receive bias voltage through the tapped primary winding. Coils *L*1 and *L*2 are RF chokes that offer a low dc resistance path for the bias current

and high ac input impedance. The circuit produces a conversion gain of 6 to 10 db, depending upon the magnitude of the local oscillator signal and the exact point of bias of the diodes.

9–15. Tunnel Diode Converters

The tunnel diode converter differs from the tunnel diode mixer (par. 9–13) in that the converter generates its own local oscillator frequency.

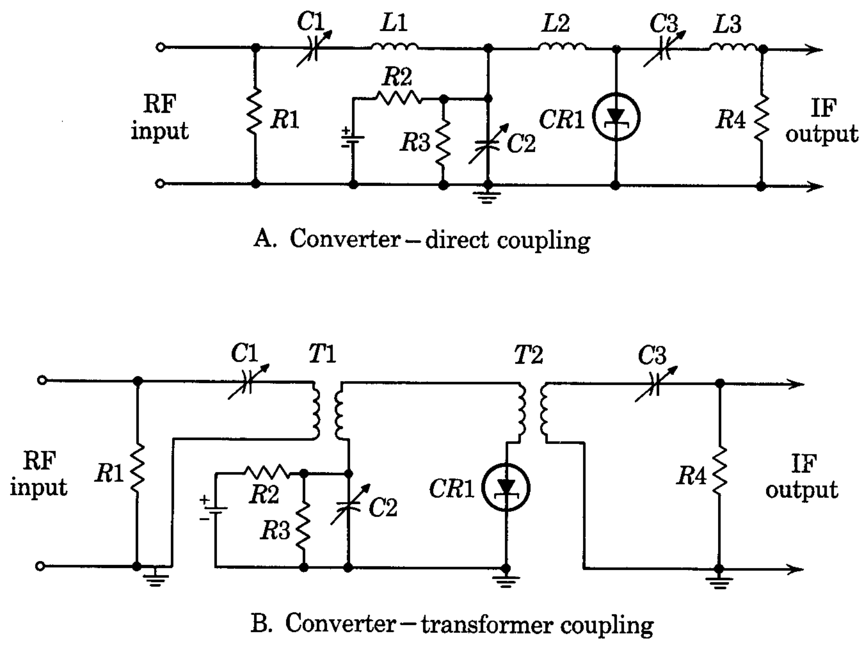


FIG. 9–16. Tunnel diode converters

A directly coupled converter is shown in Fig. 9–16A; a transformer-coupled converter is shown in Fig. 9–16B. In each circuit it is necessary to ensure oscillation at the local-oscillator frequency only. To assure this condition the circuit is heavily loaded at the RF input frequency and the IF output frequency. Heterodyne detection is obtained by virtue of the nonlinearity of the diode characteristic in its negative-resistance region.

a. Directly Coupled Converter. The converter shown in Fig. 9–16A contains three series-resonant circuits. Capacitor $C1$ and coil $L1$ are tuned to the RF input frequency and offer a low-input impedance to the RF signal only. Resistor $R1$ loads the tunnel diode at this frequency to prevent RF oscillation. Capacitor $C2$ and coil $L2$ are tuned to the desired oscillator frequency. Resistors $R2$ and $R3$ form a voltage divider to bias

diode $CR1$ in the negative-resistance region. Capacitor $C3$ and coil $L3$ are tuned to the IF frequency and offer a low-output impedance to the IF signal only. Resistor $R4$ loads the tunnel diode at the IF frequency to prevent IF oscillation. Amplification occurs at the RF and the IF frequencies so that a conversion gain of 15 to 25 db can be achieved.

b. The converter shown in Fig. 9–16B is similar to the directly coupled converter (*a* above), except that transformer coupling is used. In this case, capacitor $C1$ and the primary of transformer $T1$ are tuned to the RF signal. Capacitor $C2$, transformer $T1$ secondary, and transformer $T2$ primary are tuned to the desired oscillator frequency. Capacitor $C3$ and transformer $T2$ secondary are series tuned to the IF frequency.

9–16. Summary

a. An amplitude-modulated signal can be obtained by varying the bias of a tunnel diode oscillator in accordance with the modulating signal.

b. The controlled-negative-resistance device oscillator (par. 9–4) can be amplitude modulated by varying the control current in accordance with the modulating signal.

c. Frequency modulation of a tunnel diode oscillator depends upon the variation of the tunnel diode negative resistance which in turn helps to determine the oscillator frequency.

d. The tunnel diode AM demodulator operates as a square-law detector.

e. The tunnel diode AM demodulator can be used to detect an FM carrier by first converting the FM signal to AM by slope-tuning the input circuit.

f. Tunnel diode mixers are biased at or just below the peak-current point. The distortion thus introduced produces heterodyne action.

g. In some mixer applications it is necessary to isolate the local oscillator output from the RF amplifier. One method of achieving this condition is by use of a hybrid junction.

h. Tunnel diode converters depend upon the nonlinearity of the negative-resistance region of the diode characteristic for heterodyne detection.

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Appendix B

DERIVATION AND ANALYSIS OF FORMULAS

B–1. Series Amplifier

Figure B–1A shows a voltage source (e_s) in series with a load resistor (R_L). Figure B–1B shows the equivalent circuit of the same voltage source in series with a tunnel diode ($-R_D$) and the same load resistor.

a. Current Gain. 1. The current gain (A_i) is the ratio of the current flow i_2 (Fig. B–1B) to the current flow i_1 (Fig. B–1A):

$$A_i = \frac{i_2}{i_1}$$

2. Write the voltage equation for the circuit of Fig. B–1A:

$$e_s = i_1 R_L$$

3. Write the voltage equation for the circuit of Fig. B–1B:

$$e_s = i_2(R_L - R_D)$$

4. Equate the right-hand sides of 2 and 3:

$$i_2(R_L - R_D) = i_1 R_L$$

5. Divide both sides by $i_1(R_L - R_D)$, then:

$$A_i = \frac{i_2}{i_1} = \frac{R_L}{R_L - R_D}$$

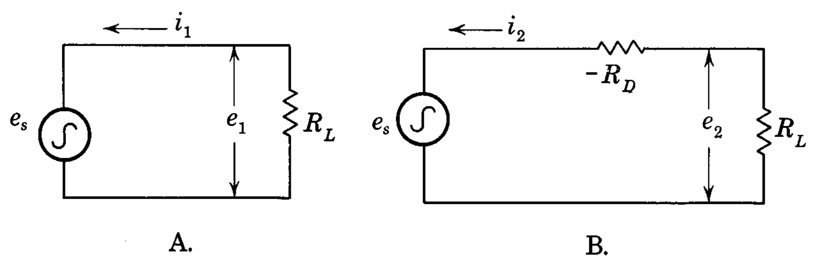


FIG. B–1. Voltage source with load and voltage source with series load and negative resistance

b. Voltage Gain. 1. The voltage gain (A_v) is the ratio of e_2 (Fig. B–1B) to e_1 (Fig. B–1A):

$$A_v = \frac{e_2}{e_1}$$

2. Note that:

$$e_2 = i_2R_L$$

and

$$e_1 = i_1R_L$$

3. Divide the two equations (2 above):

$$\frac{e_2}{e_1} = \frac{i_2}{i_1}$$

4. Note that the voltage gain (3 above) equals the current gain (*a*5 above), and therefore:

$$A_v = \frac{R_L}{R_L - R_D}$$

c. Power Gain. Power gain (G) is the product of current and voltage gain:

1. Write the expression for power gain:

$$G = A_iA_v$$

2. Substitute formulas for A_i (*a*5 above) and A_v (*b*4 above):

$$G = \frac{R_L}{(R_L - R_D)} \cdot \frac{R_L}{(R_L - R_D)}$$

3. Therefore:

$$G = \frac{R_L{}^2}{(R_L - R_D)^2}$$

B–2. Cutoff and Self-resonant Frequency

The resistance cutoff frequency (f_r) and the self-resonant frequency (f_s) of a tunnel diode are defined in paragraph 3–11. The cutoff frequency (f_r) is derived in *a* below by finding the frequency at which the positive resistance of the impedance (Z) equals the negative resistance of the impedance. Likewise the self-resonant frequency (f_s) is derived in *b* below by finding the frequency at which the positive (inductive) reactance equals the negative (capacitive) reactance.

a. Cutoff Frequency. Proceed as follows:

1. Write the impedance formula of the series-parallel equivalent circuit (Fig. B–2) of the tunnel diode:

$$Z = R_s + jX_L + \frac{(-R_D)(-jX_c)}{-R_D - jX_c}$$

Note that the vector rotator operator ($j = \sqrt{-1}$) must be used preceding reactances.

2. Consolidate negative factors in last term of above equation:

$$Z = R_s + jX_L - \frac{jX_cR_D}{R_D + jX_c}$$

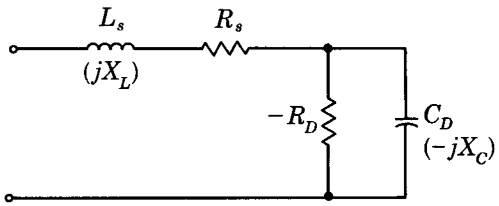


FIG. B–2. Tunnel diode equivalent circuit

3. To eliminate the vector rotator operator (j) in the denominator of the last term, multiply numerator and denominator by $(R_D - jX_c)$. (Note that $j^2 = -1$ and that the product of the sum and difference of two given terms is equal to the difference of their squares):

$$Z = R_s + jX_L - \frac{jX_cR_D}{R_D + jX_c} \cdot \frac{R_D - jX_c}{R_D - jX_c}$$

$$Z = R_s + jX_L - \frac{jX_cR_D^2 + R_DX_c^2}{R_D^2 + X_c^2}$$

4. Separate last term into resistive and inductive parts:

$$Z = R_s + jX_L - \frac{jX_cR_D^2}{R_D^2 + X_c^2} - \frac{R_DX_c^2}{R_D^2 + X_c^2}$$

5. Equate absolute values of the two resistive terms of the equation:

$$R_s = \frac{R_DX_c^2}{R_D^2 + X_c^2}$$

6. Solve for X_c:

$$X_c = R_D\sqrt{\frac{R_s}{R_D - R_s}}$$

7. Substitute for X_c:

$$X_c = \frac{1}{2\pi f_rC_D}$$

Therefore,

$$\frac{1}{2\pi f_rC_D} = R_D\sqrt{\frac{R_s}{R_D - R_s}}$$

8. Solve for f_r:

$$f_r = \frac{1}{2\pi C_DR_D}\sqrt{\frac{R_D - R_s}{R_s}}$$

Note: To use this formula, use the absolute (positive) numerical value for R_D.

b. Self-resonant Frequency. To determine the self-resonant frequency (f_s), proceed as follows:

1. Equate the positive and negative reactance terms of equation $a4$ above:

$$jX_L = \frac{jX_cR_D^2}{R_D^2 + X_c^2}$$

2. Divide both sides by j and substitute w_sL for X_L, and $\frac{1}{w_sC_D}$ for X_c, where $w_s = 2\pi f_s$:

$$w_sL = \frac{R_D^2/w_sC_D}{R_D^2 + 1/(w_sC_D)^2}$$

3. Multiply numerator and denominator by $(w_sC_D)^2$:

$$w_sL = \frac{w_sC_DR_D^2}{R_D^2(w_sC_D)^2 + 1}$$

4. Multiply both sides by the denominator:

$$(w_sL)\,(w_sC_D)^2R_D^2 + w_sL = w_sC_DR_D^2$$

5. Divide both sides by w_s:

$$w_s^2C_D^2LR_D^2 + L = C_DR_D^2$$

6. Solve for w_s^2:

$$w_s^2 = \frac{C_DR_D^2 - L}{C_D^2R_D^2L}$$

7. Take square root of both sides and substitute $2\pi f_s$ for w_s and solve for f_s:

$$f_s = \frac{1}{2\pi C_DR_D}\sqrt{\frac{C_DR_D^2 - L}{L}}$$

Note: To use this formula, use positive numerical value for R_D.

B–3. Effective Negative Resistance

a. The effective negative resistance (R_D') of the tunnel diode depends upon the shunting effect of the capacitive reactance of the inherent capacitance (C_D) of the tunnel diode (Fig. B–2). The last term of the equation of paragraph B–2*a*4 represents the value of R_D' at any operating frequency (f). Thus:

$$R_D' = \frac{R_DX_c^2}{R_D^2 + X_c^2}$$

A more usable form for this equation is derived in *b* below.

b. To transform the equation (*a* above) proceed as follows:

1. Write the reactance formula for X_c:

$$X_c = \frac{1}{2\pi fC_D}$$

2. Substitute in the equation (*a* above):

$$R_D' = \frac{R_D\left(\frac{1}{2\pi fC_D}\right)^2}{R_D^2 + \left(\frac{1}{2\pi fC_D}\right)^2}$$

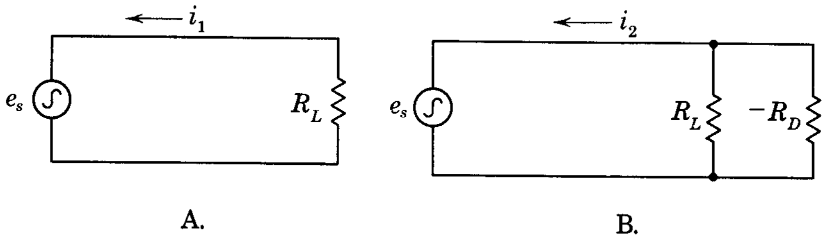


FIG. B–3. Voltage source with load and voltage source with parallel load and negative resistance

3. Multiply numerator and denominator of the complex fraction by $(2\pi f C_D)^2$:

$$R_D' = \frac{R_D}{(2\pi f C_D R_D)^2 + 1}$$

It can be readily seen from this equation that the effective negative resistance (R_D') decreases with increasing frequency. Numerical examples of the use of this formula are given in paragraph 3–18.

B–4. Parallel Amplifier

Figure B–3A shows a voltage source (e_s) in series with a load resistor (R_L). Figure 3–3B shows the equivalent circuit of the same voltage source in series with the parallel combination of the same load resistor and a tunnel diode ($-R_D$).

a. Voltage Gain. By observation it can be seen that there is no voltage gain, or:

$$A_v = \frac{e_s}{e_s} = 1$$

b. Current Gain. The magnitude of current through load resistor R_L is the same with or without the tunnel diode. However, if the current gain is defined as the ratio of *total* current (i_1) drawn from the source with load resistor R_L alone to *total* current (i_2) drawn from the source with R_L and the tunnel diode, then there is a current gain.

1. Write the current gain formula:

$$A_i = \frac{i_1}{i_2}$$

2. Write the formula for current for Fig. B–3A:

$$i_1 = \frac{e_s}{R_L}$$

3. Write the formula for current for Fig. B–3B:

$$i_2 = e_s \Big/ \frac{R_L(-R_D)}{R_L - R_D} = \frac{e_s(R_L - R_D)}{R_L(-R_D)}$$

4. Divide equations 2 and 3 above; invert the fraction in equation 3 and multiply:

$$\frac{i_1}{i_2} = \frac{e_s}{R_L} \cdot \frac{R_L(-R_D)}{e_s(R_L - R_D)}$$

5. Divide numerators and denominators by common factors:

$$A_i = \frac{i_1}{i_2} = \frac{-R_D}{R_L - R_D}$$

6. Multiply numerator and denominator by -1:

$$A_i = \frac{R_D}{R_D - R_L}$$

c. Power Gain. Power gain (G) is the product of current and voltage gain. Since voltage gain (*a* above) is unity, then:

$$G = \frac{R_D}{R_D - R_L}$$

B–5. Parallel Amplifier with Signal Source Resistance

When the load resistor (R_L) is directly in parallel with the tunnel diode (par. B–4), there is no voltage gain provided the signal source has no internal resistance. If the signal source has internal resistance, then voltage gain can be achieved as well as current gain. Derivation of applicable current, voltage, and power gain formulas for the latter case is covered in *b*, *c*, and *d* below.

a. Conversion of Voltage Generator. To simplify the derivation of the gain formulas, the voltage generator (e_s) with internal resistance (R_s) (in dashed lines in Fig. B–4A) must be converted to a current generator (i_s) with the shunt resistance (R_s) (in dashed lines in Fig. B–4B). According to network theory, a voltage generator (e_s) with internal resistance (R_s) may be replaced by a current generator having output current $i_s = e_s/R_s$ under short-circuit conditions, and an output voltage of $e_s = i_sR_s$ under open-circuit conditions. The circuits in A and B are equivalent.

b. Current Gain. Figures B–4B and C represent the same circuit, except that in the latter, the resistance values are replaced by their conductance values; furthermore the current (i_1) through and the voltage (e_1) across the load conductance are indicated in C. Figure B–4D is the equivalent circuit of C with a tunnel diode ($-G_D$) negative conductance in parallel with G_L. In Fig. B–4D, i_2 and e_2 represent the current through and the voltage across the load conductance. To derive the current gain formula, proceed as follows:

1. Write the expression for current gain:

$$A_i = \frac{i_2}{i_1}$$

2. Write the expression for i_2 (Fig. B–4D):

Note: The current through G_L is the total current multiplied by the ratio of G_L to total conductance $(G_L + G_s - G_D)$.

$$i_2 = i_s\left(\frac{G_L}{G_L + G_s - G_D}\right)$$

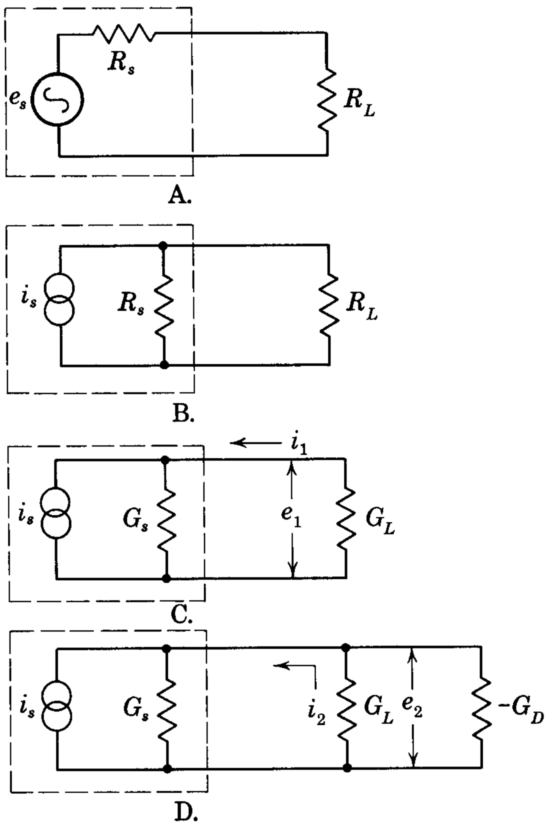


FIG. B-4. Voltage source with internal resistance and transformation of voltage source to current source with conductance values for internal resistance, load, and negative resistance

3. Write the expression for i_1 (Fig. B-4C):

$$i_1 = i_s \left(\frac{G_L}{G_L + G_s}\right)$$

4. Divide the equation in 2 above by the equation in 3 above, then:

$$A_i = \frac{i_2}{i_1} = \frac{G_L + G_s}{G_L + G_s - G_D}$$

5. Quite often it is desirable to have the current gain expressed in terms of resistance rather than conductance. To do this, substitute G_T (total positive conductance) for $G_L + G_s$, as follows:

$$A_i = \frac{G_T}{G_T - G_D}$$

6. Substitute resistance (reciprocals) for conductance values as follows:

Note: $G_T = \frac{1}{R_T}$, where R_T equals the quotient of the product and the sum (parallel value) of the two positive resistances (Fig. B–4B):

$$A_i = \frac{\frac{1}{R_T}}{\frac{1}{R_T} - \frac{1}{R_D}}$$

7. Simplify the complex fraction by multiplying numerator and denominator by $(R_T R_D)$:

$$A_i = \frac{R_D}{R_D - R_T}$$

c. Voltage Gain. To derive the voltage gain formula, proceed as follows:

1. Write the expression for voltage gain:

$$A_v = \frac{e_2}{e_1}$$

2. Note that:

$$e_2 = \frac{i_2}{G_L}$$

3. Note that:

$$e_1 = \frac{i_1}{G_L}$$

4. Divide 2 above by 3 above:

$$\frac{e_2}{e_1} = \frac{i_2}{i_1}$$

5. The voltage gain equals the current gain and:

$$A_v = \frac{R_D}{R_D - R_T}$$

d. Power Gain. To find power gain, proceed as follows:

1. Write power gain formula:

$$G = A_i A_v$$

2. Substitute expressions for A_i (*b*7 above) and A_v (*c*5 above):

$$G = \frac{R_D{}^2}{(R_D - R_T)^2}$$

B–6. Tunnel Diode Noise Figure

a. General. The noise figure of merit of the tunnel diode evaluates the noise producing characteristics of the tunnel diode alone. The *smaller* the noise figure of merit of the tunnel diode, the better the quality of the tunnel diode. The factors which affect the noise figure of merit (F) of the tunnel diode are the ac negative resistance (R_D), the series positive resistance (R_s) of the diode (par. 4–2*b*), the resistance cutoff frequency (f_r) of the diode (par. 3–17), and the operating frequency (f). One other factor is involved in the noise figure: the effect of the shot noise is represented by resis-

tance (R_e). This resistance is *inversely* proportional to the dc current through the device and the operating frequency. Considering all the factors, the noise figure of merit of the tunnel diode is expressed:

$$F = \frac{\left(1 + \frac{R_D}{R_e}\right)}{\left(1 - \frac{R_s}{R_D}\right)\left[1 - \left(\frac{f}{f_r}\right)^2\right]}$$

For the derivation of this formula see the referenced literature (Appendix A). According to this formula, the noise figure is determined by three ratios: R_D/R_e which relates the negative resistance to the shot noise; R_s/R_D which relates the thermal noise (caused by R_s) to the negative resistance; and f/f_r which relates the operating frequency to the resistance cutoff frequency. The information conveyed by this formula at low (audio) frequencies and at high frequencies is covered in *b* and *c* below.

b. Low-Frequency Considerations. At very low frequencies (including the audio frequencies) the ratio f/f_r is very small and may be considered equal to zero. The ratio R_s/R_D is approximately 0.005 in value for germanium diodes and is negligible. At low frequencies, then, the denominator of the noise figure formula (*a* above) equals *one* and the formula reduces to:

$$F = 1 + \frac{R_D}{R_e}$$

The major contribution to the noise figure at the low frequencies is due to shot noise. The equivalent shot noise resistance R_e decreases at low frequencies. As a result, at 1 kc the noise figure varies approximately from 10 to 20 db (Fig. 4–1) which is relatively high. (*Note:* db $= 10 \log_{10} F$.) At approximately 30 kc the noise figure is reduced to an average of 3 to 4 db. *This is a very low noise figure and normally better than can be achieved with electron tubes or transistors.*

c. High-Frequency Considerations. At medium frequencies [operating frequencies (f) less than half the resistance cutoff frequency (f_r)], R_D/R_e is approximately equal to unity; and R_s/R_D is approximately equal to 0.01. Note that as the operating frequency increases, the effective negative resistance R_D is reduced by the shunting effect of the inherent capacitance of the diode, thus accounting for increase in value (*b* above) of the ratio R_s/R_D. Use of these values in the noise figure formula (*a* above) will produce a noise figure close to 3 db. If f goes to $0.707\,f_r$, then F approaches 6 db and, at $f = .09 f_r$, F exceeds 10 db. The reason for the rise is that R_s/R_D approaches unity as the effective R_D is reduced, and f/f_r also approaches unity. Thus the denominator of the expression for F becomes very large. It can be stated then that *at the higher frequencies the thermal noise predominates,* and that at operating frequencies close to the resistance cutoff frequency, the noise becomes excessive.

Appendix C

REVIEW OF TRANSISTOR FUNDAMENTALS

C–1. General

Many circuits combine the desirable properties of tunnel diodes and transistors. To facilitate the discussion of these circuits, a brief review of the theory of operation of transistors is presented. The following aspects of transistor theory are discussed:

a. Current flow in *p-n-p* and *n-p-n* junction transistor (par. C–2)
b. Reference designations and graphical symbols (par. C–3)
c. General rules of transistor operation (par. C–4)
d. Transistor configurations (par. C–5)
e. Bias stabilization (par. C–6)
f. Voltage phase relationships in transistor configurations (par. C–7)
g. Comparison of configurations (par. C–8)
h. Transistor equivalent circuit and parameters (par. C–9)

C–2. Current Flow in Transistors

If a crystal containing *two p-n* junctions were prepared, signal could be introduced into one *p-n* junction biased in the forward direction (low resistance) and extracted from the other *p-n* junction biased in the reverse direction (high resistance). This would produce a *power gain* of the signal when developed in an external circuit. Such a device would transfer the signal current from a low-resistance circuit to a high-resistance circuit. Contracting the terms *transfer* and *resistor* results in the term *transistor.*

a. p-n-p Transistor. To form two *p-n* junctions, three sections of a semiconductor are required (Fig. C–1). The *p-n-p* transistor contains two sections of *p*-type semiconductor and one section of *n*-type semiconductor. When the three sections are combined, two depletion regions (barriers) occur at the junctions even though there is no application of external voltages, or fields. This phenomenon is the same as that which occurs when *two* sections of semiconductor are combined. Transistor action requires that one junction be biased in the forward direction and the second junction be biased in the reverse direction. Note that the junction between the emitter-base regions is forward biased and the junction between the base-collector regions is reverse biased. [Note that the reverse bias between the collector and the base causes a reverse-biased current (I_{CBO}) to flow. This current is caused by the minority carriers (electrons) in the collector region and the minority carriers (holes) in the base region and is the same as the reverse-biased current that occurs in a rectifying diode. This current is important in determining the temperature

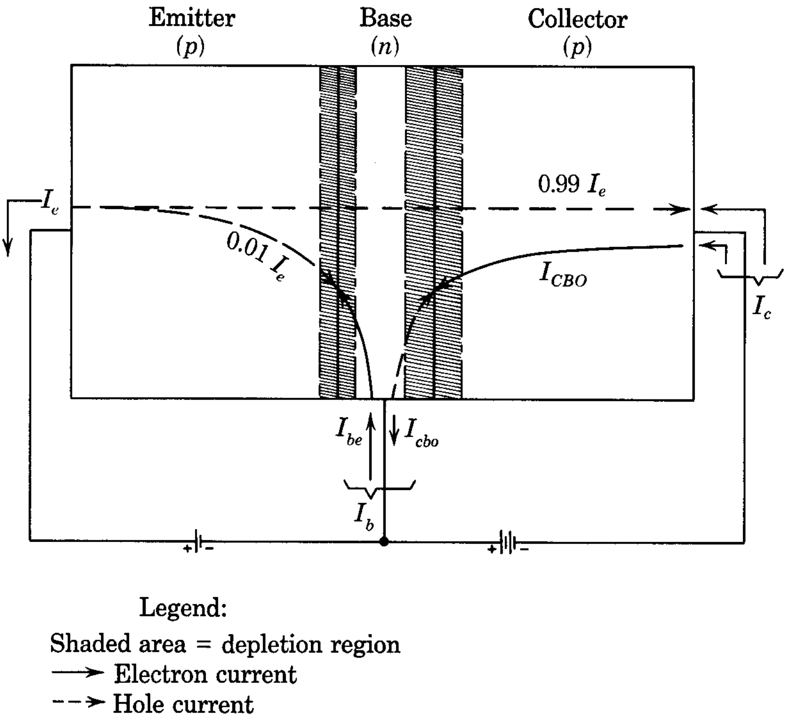


FIG. C–1. *p-n-p* transistor showing current flow and applied bias voltages

stability of the transistor (par. C–6).] Because of the simultaneous bias voltages applied, a large number of the holes from the emitter does not combine with the electrons entering the base from the emitter-base battery. About 95% to 99% of the holes diffuse through the base and penetrate the base-collector junction. In the collector region the holes combine with electrons that enter the collector from the negative terminal of the base-collector battery. If the holes that enter the base from the emitter-base junction avoid combination with electrons entering the base from the battery, the holes are attracted to the collector by the acceptor ions (negative) in the collector and the negative potential of the base-collector battery. Maximum gain is achieved if most of the emitter current (I_e) penetrates the base-collector junction. This condition is achieved in practice by making the base region as narrow as possible. The most important aspects of Fig. C–1 are as follows:

1. The end region, forward biased with respect to the base, is called the *emitter*. Its function is similar to that of the cathode in an electron tube.
2. The center region is called the *base*. Its function is similar to that of the control grid in a triode electron tube. The number of electrons that enter the base as a result of the bias (and signal) voltage determines the collector current. The greater the number of electrons, the greater the emitter and collector current.
3. The end region, reverse biased with respect to the base, is called the *collector*. Its function is similar to that of the plate in an electron tube.

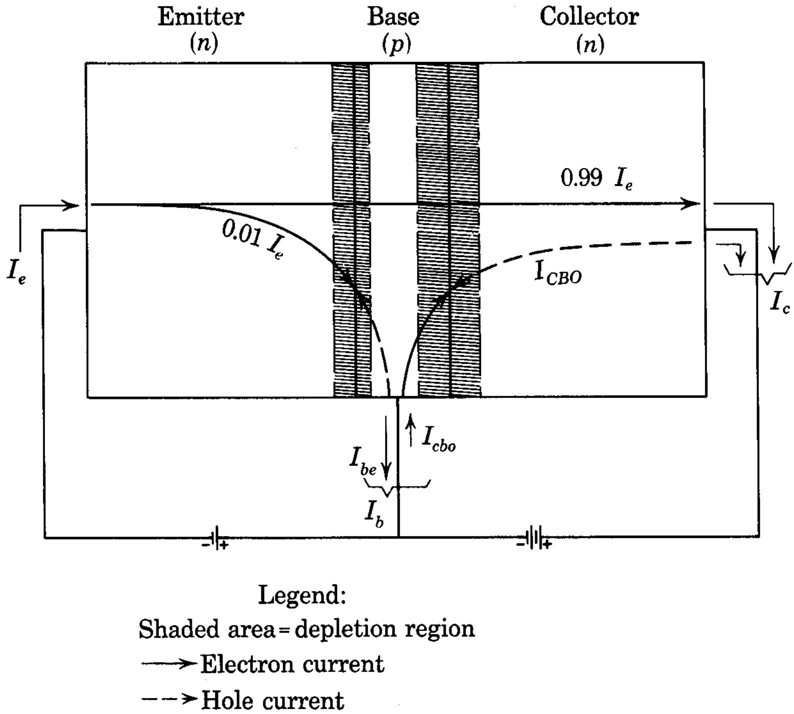


FIG. C–2. *n-p-n* transistor showing current flow and applied bias voltages

4. Note that the net base-lead current is the difference between the emitter-base current and the reverse-biased current (I_{CBO}).

b. n-p-n Transistor. The theory of operation of the *n-p-n* transistor (Fig. C–2) is similar to that of the *p-n-p* transistor (*a* above). However, there are two important differences:

1. The emitter-to-collector carrier in the *p-n-p* transistor is the hole. The emitter-to-collector carrier in the *n-p-n* transistor is the electron.
2. The bias voltage polarities are reversed because of the different positional relationships of the two types of semiconductor materials used.

C–3. Reference Designations and Graphical Symbols

The reference designations and graphical symbols for transistors are shown in Figs. C–3A and B.

a. Reference Designations. The letter portion of the transistor reference designation is *Q* and the number portion may be any whole number.

b. Graphical Symbols. 1. In the *p-n-p* transistor (Fig. C–3A) the emitter-to-collector current carrier in the crystal is the hole (par. C–2). For holes to flow internally from emitter to collector, the collector must be negative with respect to the emitter. In the external circuit, electrons flow from the emitter (*opposite* to direction of the emitter arrow) to the collector.

2. In the *n-p-n* transistor (Fig. C–3B), the emitter-to-collector current carrier in the crystal is the electron (par. C–2). For electrons to flow internally from emitter to collector, the collector must be positive with respect to the emitter. In the external circuit, electrons flow from the collector to the emitter (opposite to direction of the emitter arrow).

C–4. Transistor Amplifier General Rules

The following general rules will help in the qualitative (nonmathematical) analysis of transistor circuits. These general rules apply to a *class A* transistor amplifier.

a. The external dc electron current direction is always *against* the direction of the arrow on the emitter (Figs. C–3A and B).

b. If the electrons flow into or out from the emitter, the electrons flow out from or into the collector, respectively.

c. The base-emitter junction is always *forward* biased.

d. The collector-base junction is always *reverse* biased.

e. An input voltage that *aids* (increases) the forward bias *increases* the emitter and collector currents.

f. An input voltage that *opposes* (decreases) the forward bias *decreases* the emitter and collector currents.

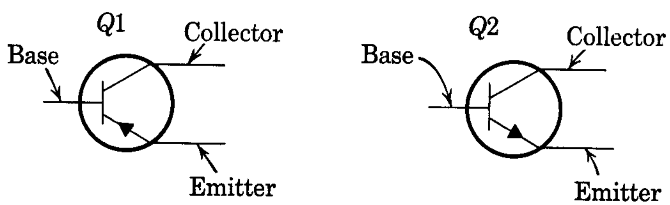
g. The first letter of the transistor type indicates the emitter polarity with respect to the base. The *p-n-p* transistor emitter is at a positive dc voltage. The *n-p-n* transistor emitter is at a negative dc voltage.

h. The second letter of the transistor type indicates the collector polarity with respect to the base. The *p-n-p* transistor collector is at a negative dc voltage. The *n-p-n* transistor collector is at a positive dc voltage.

i. The first and second letter of the transistor type indicate the relative polarities between the emitter and the collector, respectively. The *p-n-p* transistor emitter is positive with respect to the collector; the collector is negative with respect to the emitter. The *n-p-n* transistor emitter is negative with respect to the collector; the collector is positive with respect to the emitter.

C–5. Transistor Configurations

The transistor may be connected in a common base, common emitter, or a common collector configuration. The designation depends on which element is common to the input and the output circuit. In electronic equipments, inasmuch as the



A. *p-n-p* transistor B. *n-p-n* transistor

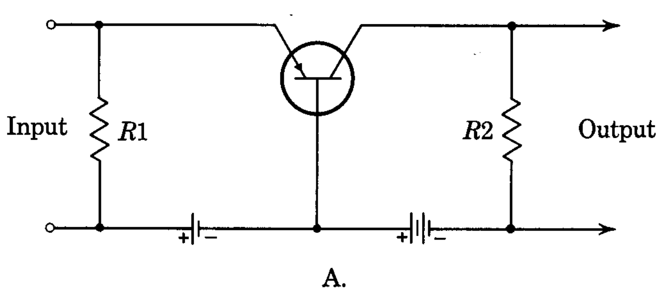
FIG. C–3. Transistor reference designations and graphical symbols

common element is most often at ac ground potential or ac and dc ground (chassis) potential, these configurations are also referred to as grounded base, grounded emitter, or grounded collector. Regardless of the configuration used, the class A amplifier requires forward bias of the base-emitter junction and reverse bias of the collector-base junction.

Note: This paragraph discusses configurations using *p-n-p* transistors; the discussion also applies to *n-p-n* transistors if the bias batteries are reversed in polarity.

a. Common-Base (CB) Amplifier. Figure C–4 shows the common-base configuration in which the signal is introduced into the emitter-base circuit, and extracted from the collector-base circuit. The base element of the transistor is common to the input circuit and the output circuit. Figure C–4A shows the use of two separate batteries for biasing; one battery forward biases the base-emitter junction, and a second battery forward biases the collector-base junction. Figure C–4B shows the use of one battery for collector-base and emitter-base biasing. Forward bias is achieved by using the voltage divider consisting of resistors $R3$ and $R4$.

b. Common-Emitter (CE) Amplifier. Figure C–5 shows the common emitter configuration in which the signal is introduced into the base-emitter circuit and extracted from the collector-emitter circuit. The emitter element of the transistor is common to the input circuit and the output circuit. Figure C–5A shows the use of two biasing batteries: one to forward bias the base-emitter junction, and the



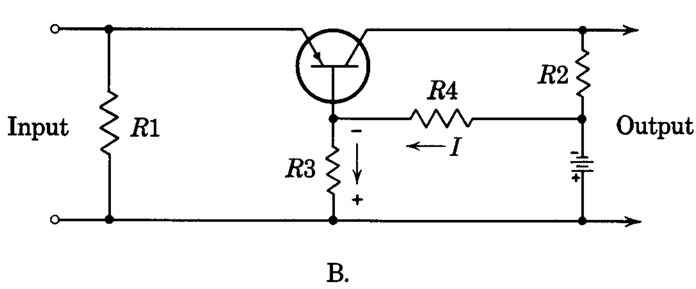


FIG. C–4. Common base configurations, using two-battery and single-battery biasing

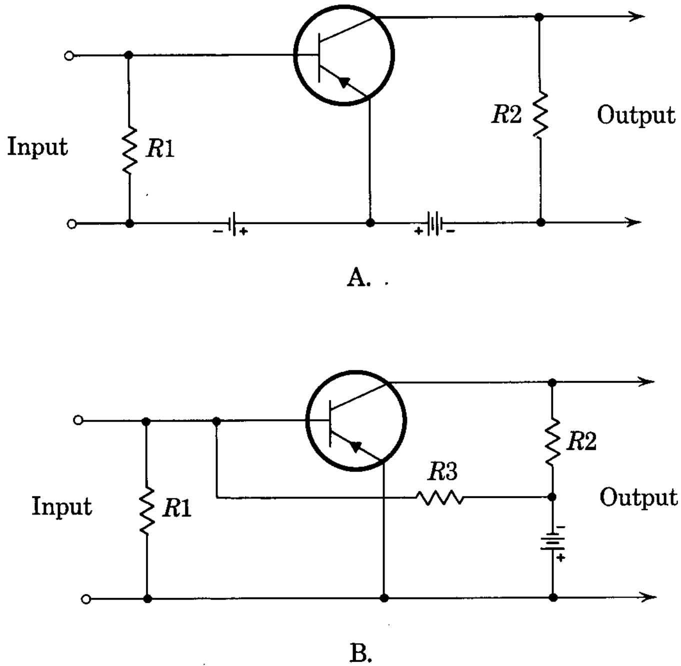


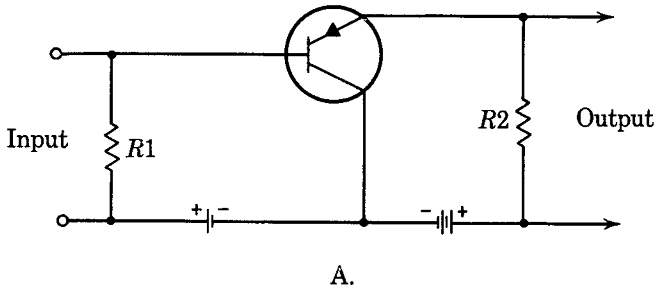
FIG. C–5. Common emitter configuration using two-battery and single-battery biasing

second to reverse bias the collector-base junction. Note that the battery in the emitter-collector circuit is larger than the battery in the emitter-base circuit. Figure C–5B shows the use of one battery to bias the base-emitter junction and the base-collector junction. Forward bias is achieved by using the voltage divider consisting of resistors $R3$ and $R1$.

c. Common-Collector (CC) Amplifier. Figure C–6 shows the common collector configuration in which the signal is introduced into the base-collector circuit and extracted from the emitter-collector circuit. The collector element of the transistor is common to the input circuit and the output circuit. Figure C–6A shows the use of two batteries to obtain the proper biases; their potentials are such that the base-emitter junction is forward biased and the collector-base is reverse biased. Figure C–6B shows the use of one battery for biasing. To obtain forward bias in the base-emitter circuit, a voltage divider consisting of resistors $R1$ and $R3$ is used.

C–6 Bias Stabilization

Bias (operating point) is established for a transistor by specifying the quiescent (dc, no-signal) values of output voltage and input current. Distortionless operation of a transistor over a wide range of temperatures requires that bias voltage and current remain stable. However, variations of reverse-biased collector current (*a* below) and emitter-base junction resistance (*b* below) with temperature pre-



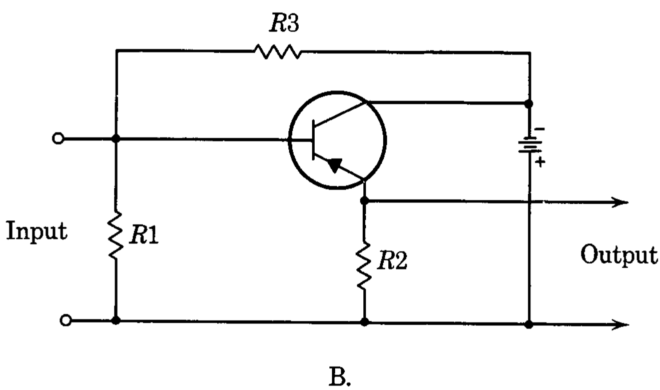


FIG. C–6. Common collector configuration using two-battery and single-battery biasing

clude stable bias unless external compensating circuits are employed. Bias stabilization using resistors is discussed in *c* below.

a. Reverse-bias Collector Current (I_{CBO}). 1. The reverse-biased collector current (par. C–2) is also referred to as a leakage current (because it is undesirable) and *saturation* current (because it reaches its maximum value at a given temperature with a small amount of reverse bias). This current is designated I_{CBO} because it is the *c*ollector-*b*ase current measured with the emitter *o*pen. The saturation current increases when the temperature of the transistor increases. The saturation current value varies from almost zero at 10° C to well over 1 milliampere at 125° C.

2. Figures C–1 and C–2 show the direction of current flow in the base lead of a *p-n-p* and an *n-p-n* transistor caused by the saturation current. In each case the direction of current is such that a *forward bias* in the base-emitter circuit will be created by this current if resistance is present in the base lead. This increased forward bias increases the collector current. Increased collector current would raise the temperature of the collector-base junction and increase the saturation current. The cycle would continue until distortion occurs; the transistor becomes inoperative, or it destroys itself. *This condition can be minimized by avoiding the use of high-valued resistors in the base lead.*

b. Emitter-Base Junction Resistance. 1. If the collector current variation with temperature were caused only by the saturation current, then collector current variation at temperatures below 10° C should not occur (*a*1 above). However, collector current increases with temperature even when the saturation current is near zero. This variation is caused by the decrease in emitter-base junction resistance when the temperature is *increased;* i.e., the emitter-base junction resistance has a *negative temperature coefficient of resistance.*

2. One method of reducing the effect of the negative temperature coefficient of resistance is to place a large-valued resistor in the *emitter* lead. This causes the variation of the emitter-base junction resistance to be a small percentage of the *total* resistance in the emitter circuit. The external resistor *swamps* (overcomes) the junction resistance; the resistor is referred to as a *swamping resistor* and is usually bypassed by a capacitor to avoid negative feedback of the ac signal.

3. A second method of reducing the effect of the negative temperature coefficient of resistance is to reduce the emitter-base forward bias as the temperature increases. This method employs temperature-sensitive elements such as thermistors and junction diodes in the biasing circuits of the transistor amplifier.

c. Bias Stabilization Using Resistors. 1. The common base configuration shown in Figure C–4A has good temperature stability because the base-lead resistance is zero and emitter load resistor *R*1 also acts as an emitter swamping resistor.

2. The common collector configurations shown in Fig. C–6 would have good temperature stability if base resistor *R*1 were replaced by a low-resistance coil or if transformer coupling were used to minimize base resistance. Emitter load resistor *R*2 also acts as an emitter swamping resistor.

3. One method of obtaining temperature stability in a *CE* amplifier uses near-zero base resistance and an emitter swamping resistor as shown in Fig. C–7A. Resistor *R*1, ac bypassed by capacitor *C*1, is the swamping resistor; the secondary of transformer *T*1 offers a very low dc resistance in the base circuit. Resistor *R*2 is the collector load. The collector current for this circuit will remain stable over a wide operating temperature.

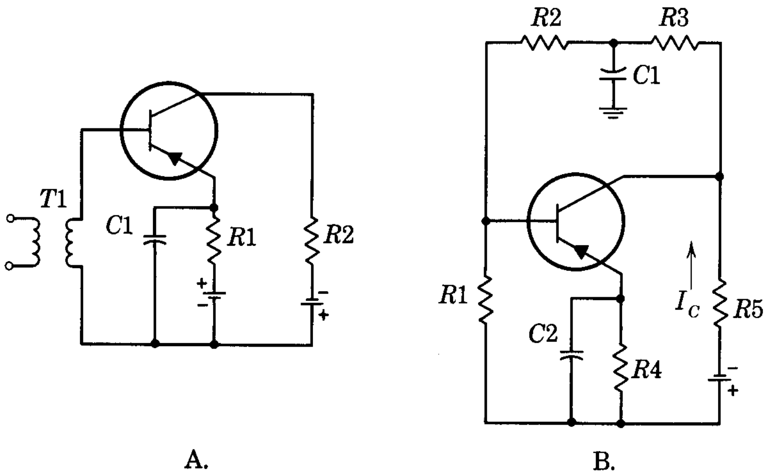


FIG. C–7. *CE* amplifiers employing transformer input and self-bias

4. A circuit employing negative feedback voltage to improve current stability is shown in Fig. C–7B. This method of bias is referred to as self-bias. If the collector current (I_C) rises, the collector becomes less negative because of the larger dc drop in resistor *R*5. As a result, less forward bias (negative base to positive emitter) is coupled through resistors *R*2 and *R*3 to the base. Reduced forward bias then reduces the collector current. Resistor *R*2 isolates the base terminal from ac decoupling capacitor *C*1; resistor *R*3 isolates the collector terminal from ac decoupling capacitor *C*1. Resistor *R*4 is the emitter swamping resistor, ac bypassed by capacitor *C*2. Resistor *R*5 is the collector load and develops the output signal.

C–7. Voltage Phase Relations

Electron current flow through *CB*, *CE*, and *CC* amplifiers using *p-n-p* transistors is indicated by the direction of the arrows in Figs. C–8, C–9, and C–10 respectively. To simplify the explanation of the circuits, the saturation current has been ignored. Most of the emitter current flows through the collector. In practical transistors, from 95 to 99% of the emitter current reaches the collector; the remainder flows through the base. In the figures, the total emitter current is represented by the letter *I*. For discussion purposes, it is assumed that 99% (or 0.99 *I*) of the current reaches the collector; 1% (or 0.01 *I*) flows to the base. The waveforms on the illustrations represent *voltage* waveforms. The input signal produced by the signal generator is on the left and the output signal developed across resistor *R*1 is on the right. If *n-p-n* transistors were to be used, the polarities of the batteries would have to be reversed, but the voltage phase relationships would be the same.

a. CB Amplifier (Fig. C–8). 1. Consider an instant of time when the voltage (*AB*) from the signal generator aids the base-emitter forward bias. The resultant forward bias at this instant has been increased, thereby increasing the total emitter current (*I*). By corresponding amounts, the collector and the base currents have been increased. The increased current flow through resistor *R*1 causes the top part of the resistor to become more positive with respect to the lower part. This effect is shown by *AB* on the output waveform. For the entire half cycle that the input signal *goes positive* and aids the forward bias, the output signal goes positive.

2. Consider an instant of time when the voltage (*CD*) from the input signal opposes the forward bias. The decreased forward bias decreases the total emitter current (*I*). By corresponding amounts, the collector and the base currents have been decreased. The decreased current flow through resistor *R*1 causes the top part of the resistor to become less positive with respect to the lower part. This effect is shown by *CD* on the output waveform. For the entire half cycle that the

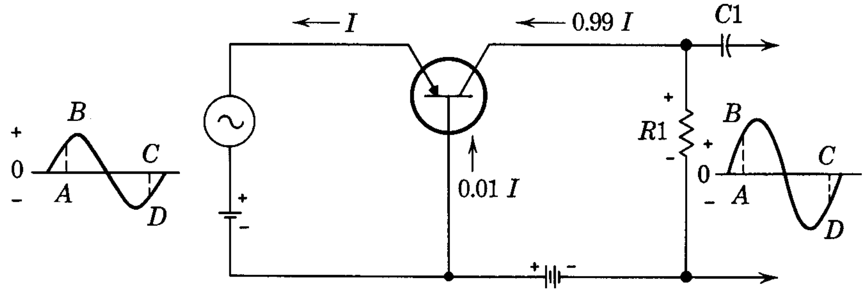


FIG. C–8. Common base (*CB*) amplifier, input and output voltage waveforms

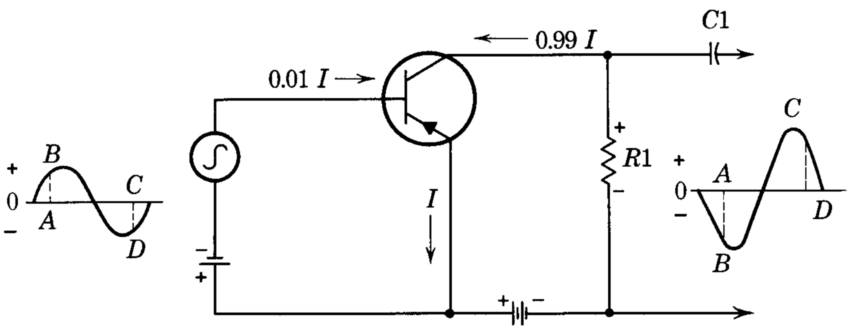


FIG. C–9. Common-emitter (CE) amplifier, input and output voltage waveforms

input signal *goes negative* and *opposes* the forward bias, the output signal goes negative.

3. Comparing the input and output signals, therefore, it can be seen that there is no voltage phase reversal in a CB amplifier.

b. CE Amplifier (Fig. C–9). 1. Consider an instant of time when the input voltage (AB) opposes the forward bias of the base-emitter battery. The reduced forward voltage at this instant decreases the total emitter current (I). By corresponding amounts the collector and the base currents are decreased. The decreased current through resistor $R1$ causes the top part of the resistor to become less positive with respect to the lower part. This effect is shown by AB on the output waveform. For the entire half cycle that the input signal *goes positive* and opposes the forward bias, the output signal goes negative.

2. Consider an instant of time when the input voltage (CD) aids the forward bias. The increased forward voltage increases the emitter current. By corresponding amounts the collector and the base currents are increased. The increased current through resistor $R1$ causes the top part of the resistor to become more positive with respect to the lower part. This effect is shown by CD on the output waveform. For the entire half cycle that the input signal *goes negative* and aids the forward bias, the output signal goes positive.

3. Comparing the input and the output signals, therefore, it can be seen that the input signal voltage is reversed 180° in phase in going through the CE amplifier.

c. CC Amplifier (Fig. C–10). 1. Consider an instant of time when the input voltage (AB) opposes the forward-bias voltage. The reduced forward voltage decreases the emitter current (I). The decreased current through resistor $R1$ causes the top part of the resistor to become less negative (more positive) with respect to the lower part. This effect is shown by AB on the output waveform. For the entire half cycle that the input signal goes positive, the output signal also goes positive.

2. When the input signal is negative (CD), the forward bias is increased and the emitter current increases. The increase in emitter current causes the top part of resistor $R1$ to become more negative with respect to the lower part. This effect is shown by CD on the output waveform. For the entire half cycle that the input signal goes negative, the output signal also goes negative.

3. Comparing the input and output signals, it can be seen that there is no phase reversal of the signal by the CC amplifier.

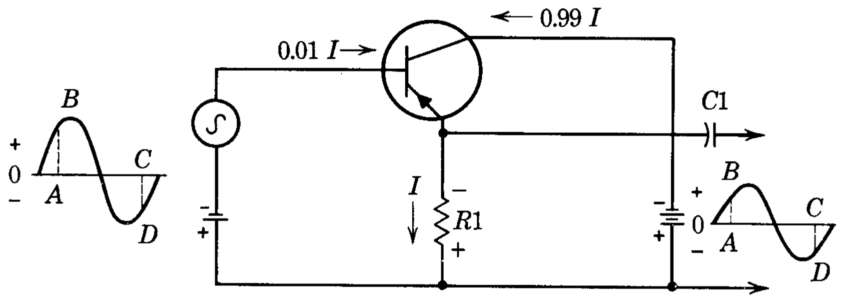


FIG. C–10. Common-collector (*CC*) amplifier, input and output voltage waveforms

C–8. Comparison of Configurations

a. General. The three possible circuit configurations for the transistor were covered in paragraph C–5. Typical values of input resistance, output resistance, current gain, voltage gain, and power gain for medium-power transistors are given in *b* below. This information will help the reader to appreciate the combinations of configurations used in specific applications, and the combinations of configurations used to achieve a given result. In equipments containing transistors, the *CE* amplifier is most often used because it provides current, voltage, and power gain, while the *CB* amplifier current gain is always less than 1 and the *CC* amplifier voltage gain is always less than 1. The *CB* amplifier is used most often to match a low-impedance circuit to a high-impedance circuit. The *CC* amplifier is used most often to match a high-impedance circuit to a low-impedance circuit.

b. Typical Values. Table C–1 lists typical values of input resistance, output resistance, voltage gain, current gain, and power gain for the three configurations.

TABLE C–1. TYPICAL CONFIGURATION VALUES

Item	*CB* Amplifier	*CE* Amplifier	*CC* Amplifier
Input resistance	30–150 ohms	500–1500 ohms	2K–500K ohms
Output resistance	300K–500K	30K–50K	50–1000 ohms
Voltage gain	500–1500	300–1000	Less than 1
Current gain	Less than 1	25–50	25–50
Power gain	20–30 db	25–40 db	10–20 eb

C–9. Transistor Equivalent Circuit and Parameters

Figure C–11 shows a *CE* amplifier and an equivalent circuit by which the amplifier can be represented. The transistor (in dashed lines) is represented by a resistor h_{ie}, a voltage generator with an output voltage $h_{re}v_{ce}$, a resistor having a conductance h_{oe}, and a current generator having an output current $h_{fe}i_b$. These parameters are used extensively in the field and are referred to as *hybrid* parameters. They are so called because they represent a *mixture* of dimensions; i.e., h_{ie} is meas-

TABLE C–2. HYBRID PARAMETER VALUES

Common Emitter		Common Base		Common Collector	
Parameter	Typical Value	Parameter	Typical Value	Parameter	Typical Value
$h_{ie} = \left(\frac{\Delta V_{BE}}{\Delta I_B}\right)_{V_{CE}=C}$	1950 ohms	$h_{ib} = \left(\frac{\Delta V_{EB}}{\Delta I_E}\right)_{V_{CB}=C}$	39 ohms	$h_{ic} = \left(\frac{\Delta V_{BC}}{\Delta I_B}\right)_{V_{EC}=C}$	1950 ohms
$h_{re} = \left(\frac{\Delta V_{BE}}{\Delta V_{CE}}\right)_{I_B=C}$	575×10^{-6}	$h_{rb} = \left(\frac{\Delta V_{EB}}{\Delta V_{CB}}\right)_{I_E=C}$	380×10^{-6}	$h_{rc} = \left(\frac{\Delta V_{BC}}{\Delta V_{EC}}\right)_{I_B=C}$	1
$h_{fe} = \left(\frac{\Delta I_C}{\Delta I_B}\right)_{V_{CE}=C}$	49	$h_{fb} = \left(\frac{\Delta I_C}{\Delta I_E}\right)_{V_{CB}=C}$	−0.98	$h_{fc} = \left(\frac{\Delta I_E}{\Delta I_B}\right)_{V_{EC}=C}$	−50
$h_{oe} = \left(\frac{\Delta I_C}{\Delta V_{CE}}\right)_{I_B=C}$	24.5 μmhos	$h_{ob} = \left(\frac{\Delta I_C}{\Delta V_{CB}}\right)_{I_E=C}$	0.49 μmhos	$h_{oc} = \left(\frac{\Delta I_E}{\Delta V_{EC}}\right)_{I_B=C}$	24.5 μmhos

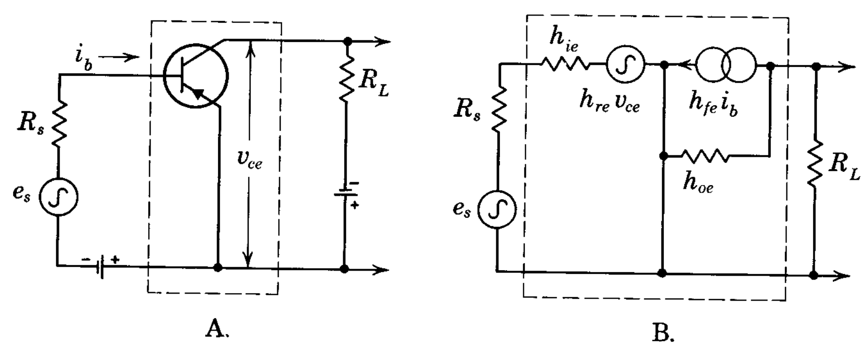


FIG. C–11. *CE* amplifier and its equivalent circuit

ured in ohms and h_{oe} in mhos, while h_{fe} and h_{re} are pure numbers. Other equivalent circuits can be derived using all *resistance* values or all *conductance* values. The hybrid parameters are defined in *a* below, and specific formulas and typical values are given in *b* below.

a. Definition. The general definitions for each parameter (1 through 4 below) applies to the *CE*, *CB*, or *CC* configuration; each is made applicable to the specific configuration by adding *e*, *b*, or *c*, respectively, to the subscript.

1. h_i is the input resistance measured with the output ac short-circuited.
2. h_r is the reverse (feedback) open-circuit voltage amplification factor and is measured with the input ac open-circuited.
3. h_f is the forward current amplification factor and is measured with the output ac short-circuited.
4. h_o is the output conductance measured with the input ac open-circuited.

b. Formulas and Typical Values. Table C–2 lists the hybrid parameters for the *CE*, *CB*, and *CC* configurations. In the case of voltage (V) and current (I), the subscripts C, B, E, refer to collector, base, and emitter, respectively; where C is used alone, it means *constant*. For instance, the first formula reads: the input resistance (h_{ie}) equals the ratio of the incremental (Δ) values of base-emitter voltage (V_{BE}) to base current (I_B), measured with the collector-emitter voltage (V_{CE}) constant (C). The table also presents a set of values for the hybrid parameters of a typical transistor. Actually, the parameter values for a given configuration vary over a very wide range, depending on the transistor types. The value of the table, therefore, is in showing the relative magnitudes in parameter values among the three configurations.

GLOSSARY

Acceptor Impurity. A substance with three (3) valence electrons in its atom; added to a semiconductor crystal, it creates a positive mobile hole in the lattice structure of the crystal.

AND Circuit (AND Gate). A gating circuit that switches from one stable state to another when all inputs are applied simultaneously.

Astable Multivibrator. A multivibrator biased in an unstable region and switching rapidly and continuously through the unstable region when power is applied; referred to as free running.

Backward Diode. A semiconductor device in which the tunneling current at a given small reverse bias is *greater* than the tunneling current at the same forward bias.

Barrier. In a semiconductor, the electric field between the acceptor ions and the donor ions at a junction. See **Depletion Region.**

Barrier Height. In a semiconductor, the difference in potential from one side of a barrier to the other.

Base (junction transistor). The center semiconductor region of a double junction (*n-p-n* or *p-n-p*) transistor. The base is comparable to the control grid of a triode electron tube.

Bistable Multivibrator. A circuit with two stable states requiring two input pulses to complete a cycle.

Collector. The end semiconductor material of a double junction (*n-p-n* or *p-n-p*) transistor that is normally reverse-biased with respect to the base. The collector is comparable to the plate of an electron tube.

Common-Base (*CB*) Amplifier. A transistor amplifier in which the base element is common to the input and the output circuit. This configuration is comparable to the grounded-grid triode electron tube.

Common-Collector (*CC*) Amplifier. A transistor amplifier in which the collector element is common to the input and the output circuit. This configuration is comparable to the electron-tube cathode follower.

Common-Emitter (*CE*) Amplifier. A transistor amplifier in which the emitter element is common to the input and the output circuit. This configuration is comparable to the conventional electron-tube amplifier.

Conductance, Negative. See **Negative Conductance.**

Configuration. The relative arrangement of parts (or components) in a circuit.

Cutoff Frequency, Resistance. See **Resistance Cutoff Frequency.**

Depletion Region (or Layer). The region in a semiconductor containing the uncompensated acceptor and donor ions; also referred to as the space-charge region or barrier region.

Donor Impurity. A substance with five (5) electrons in the outer orbit of its atom; added to a semiconductor crystal, it provides one free electron.

Dwell Period. That period during which a switching device remains for a relatively long time at a given operating (voltage or current) level; the time between two switching periods.

Electron-Pair Bond. A valence bond formed by two electrons, one from each of two adjacent atoms.

Emitter-Follower Amplifier. See **Common-Collector Amplifier.**

Emitter (junction transistor). The end semiconductor material of a double junction (*p-n-p* or *n-p-n*) transistor that is normally forward-biased with respect to the base. The emitter is comparable to the cathode of an electron tube.

Equivalent Circuit. A diagrammatic circuit representation of any device exhibiting two or more electrical parameters.

EXCLUSIVE OR Gate. A circuit which may have a high-voltage output *or* a low-voltage output depending upon the magnitude of the input pulses.

Fall Time. The length of time during which the amplitude of a pulse decreases from 90% to 10% of its maximum value.

Forward Bias. An external potential applied to a *p-n* junction so that the depletion region is narrowed and relatively high current flows through the junction.

Forward Short-Circuit Current Amplification Factor. In a transistor, the ratio of incremental values of output to input current when the output circuit is ac short-circuited.

Forward Voltage. See **Injection Current Voltage.**

Gating Circuit. A circuit operating as a switch, making use of a short or open circuit to apply or eliminate a signal.

Hall Effect. The electromotive force (voltage gradient) developed in a current-carrying conductor when immersed in a magnetic field. The direction of the electromotive force is at right angles to the direction of current flow and the magnetic field.

Hole. A mobile vacancy in the lattice structure of a semiconductor crystal. The hole acts similarly to a positive electronic charge having a positive mass.

Hybrid Junction. As used in this text, a four-port hollow waveguide device that permits electrical energy flow from one port to only *two* other ports; used as an isolating device.

Hybrid Parameter. The parameters of an equivalent circuit of a transistor which are not dimensionally identical.

Inflection Point. That point on the current-voltage (I-V) chart of a tunnel diode that displays minimum negative resistance (maximum negative conductance).

Injection Current. Current flow across a forward-biased semiconductor junction that occurs because of the high energy levels of the mobile charges. In a tunnel diode it is the current flow beyond the valley current.

Injection Current Voltage. In a tunnel diode, the applied *forward voltage* that causes an injection current equal to the peak current.

Junction Transistor. A device having three alternate sections of *p*-type and *n*-type semiconductor material.

Lattice Structure. In a crystal, a stable arrangement of atoms and their electron-pair bonds.

Majority Carriers. The holes or free electrons in *p*-type or *n*-type semiconductors, respectively.

MAJORITY Gate. A circuit, the output signal magnitude or polarity of which depends upon the polarity of the majority of the inputs.

Maser. Word derived from *M*icrowave *A*mplification by *S*timulated *E*mission of *R*adiation. Term refers to process of, or device for, microwave amplification or oscillation by direct interaction of electromagnetic fields and the molecular structure of a device.

Minority Carriers. The holes or excess electrons in *n*-type or *p*-type semiconductors, respectively.

Monostable Multivibrator. A circuit having one stable state and requiring one trigger pulse to complete a cycle.

Multivibrator. A type of relaxation oscillator used to generate nonsinusoidal waves.

Nanosecond. One billionth of a second; one millimicrosecond; 10^{-9} second.

Negative Conductance. The reciprocal of negative resistance.

Negative Resistance. The *ac* quantitative (mathematical) effect of a device the instantaneous dc positive resistance value of which *increases* with increasing voltage (i.e., current decreases with increasing voltage).

NOR Circuit. An *OR* gating circuit that provides pulse phase inversion.

NOT AND Circuit. An *AND* gating circuit that provides pulse phase inversion.

NOT Gate. A phase inverter; the output pulse is reversed 180° in phase with respect to the input pulse.

***n*-Type Semiconductor.** A semiconductor into which a donor impurity has been introduced. It contains free electrons.

OFF State. With respect to the tunnel diode, the device is biased above the valley voltage.

ON State. With respect to the tunnel diode, the device is biased below the peak voltage.

OR Circuit (OR Gate). A gate circuit that produces the desired output with only one of several possible input signals applied.

Peak Current. The maximum current that occurs in a forward-biased tunnel diode during the tunneling process.

Peak-Current Voltage. The tunnel diode voltage at which the peak current occurs.

***p-n* Junction.** The area of contact between *n*-type and *p*-type semiconductor materials that are lightly doped.

p_T-*n* Junction. A *p-n* junction formed by heavily doped semiconductor regions and resulting in electron *t*unneling.

Polycrystalline Structure. The granular structure of crystals which are nonuniform in shape and irregularly arranged.

Precession. The rotary motion of the *axis* of spin of a rotating body when a force is applied at an angle to the axis of spin. If the body were not rotating (did not have angular momentum), the body would simply move in the direction of the applied force. The rotary motion of the axis of spin, therefore, takes *precedence* over the latter possible motion.

***p*-Type Semiconductor.** A semiconductor crystal into which an acceptor impurity has been introduced. It provides holes in the crystal lattice structure.

Pulse Time. The length of time a pulse remains at its maximum value.

Quiescence. The operating condition that exists in a circuit when no input signal is applied to the circuit.

Resistance Cutoff Frequency. The frequency above which a negative-resistance device cannot oscillate because its total positive resistance equals its negative

resistance. The frequency at which the real part of its impedance equation equals zero.

Resistance, Negative. See **Negative Resistance.**

Reverse Bias. An external potential applied to a *p-n* junction that increases the junction barrier and prevents the movement of majority current carriers.

Rise Time. The time required for the leading edge of a pulse to increase from 10% to 90% of its maximum value.

Semiconductor. A substance having a resistivity between that of conductors and insulators.

Stabilization. The reduction of variations in voltage or current not due to prescribed conditions.

Swamping Resistor. In transistor circuits, a resistor placed in the emitter lead to mask (or minimize the effects of) variations in emitter-base junction resistance caused by variations in temperature.

Switching Period. That period during which a switching device moves almost instantaneously from one operating (current or voltage) level to another.

Thermal Agitation. In a semiconductor, the random movement of holes and electrons within a crystal due to the thermal (heat) energy.

Transistor. A semiconductor device capable of transferring a signal from one circuit to another and producing amplification. See **Junction Transistor.**

Triggered Circuit. A circuit that requires an input signal (trigger) to produce a desired output determined by the characteristics of the circuit.

Tunnel Diode. A heavily doped semiconductor diode that exhibits a negative-resistance region in its current-voltage chart because of electron tunneling.

Tunneling. The movement of electrons across a normally impassable narrow barrier when holes (or other positive ions) exist on the other side of the barrier at the same energy level as the electron.

Tunneltron. A device consisting of a sandwich of two conductors and an ultra-thin insulator and exhibiting the electron tunneling phenomenon.

Unilateralization. The process by which an amplifier or cascaded amplifiers are prevented from oscillating by ensuring the unidirectional flow of energy through the circuit.

Unit Step Current (or Voltage). A current (or voltage) that undergoes an instantaneous change in magnitude from one constant level to another.

Valley Current. The minimum current that occurs in a tunnel diode forward biased beyond the peak voltage. The transition point between tunneling current and injection current.

Valley Current Voltage. The tunnel diode voltage at which the valley current occurs.

Zener Diode. A *p-n* junction diode reverse-biased into the breakdown region; used for voltage stabilization.

INDEX

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